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(12) **United States Patent**
Khlat et al.

(10) **Patent No.: US 10,141,644 B2**
(45) **Date of Patent: Nov. 27, 2018**

(54) **ACOUSTIC FILTER FOR ANTENNAS**

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(73) Assignee: **Qorvo US, Inc.**, Greensboro, NC (US)

(*) Notice: Subject to any disclaimer, the term of this patent is extended or adjusted under 35 U.S.C. 154(b) by 0 days.

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(22) Filed: **Apr. 18, 2017**

(65) **Prior Publication Data**

US 2017/0301992 A1 Oct. 19, 2017

Related U.S. Application Data

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(51) **Int. Cl.**

H03H 9/74	(2006.01)
H01Q 5/335	(2015.01)
H03H 9/64	(2006.01)
H03H 9/60	(2006.01)
H04B 1/525	(2015.01)

(Continued)

(52) **U.S. Cl.**
CPC ***H01Q 5/335*** (2015.01); ***H03H 9/542*** (2013.01); ***H03H 9/605*** (2013.01); ***H03H 9/6483*** (2013.01); ***H03H 9/703*** (2013.01); ***H03H 9/72*** (2013.01); ***H04B 1/0064*** (2013.01); ***H04B 1/525*** (2013.01); ***H04B 11/00*** (2013.01); *H01Q 1/242* (2013.01); *H03H 7/09* (2013.01)

(58) **Field of Classification Search**
CPC H01Q 5/335
USPC 343/853, 858
See application file for complete search history.

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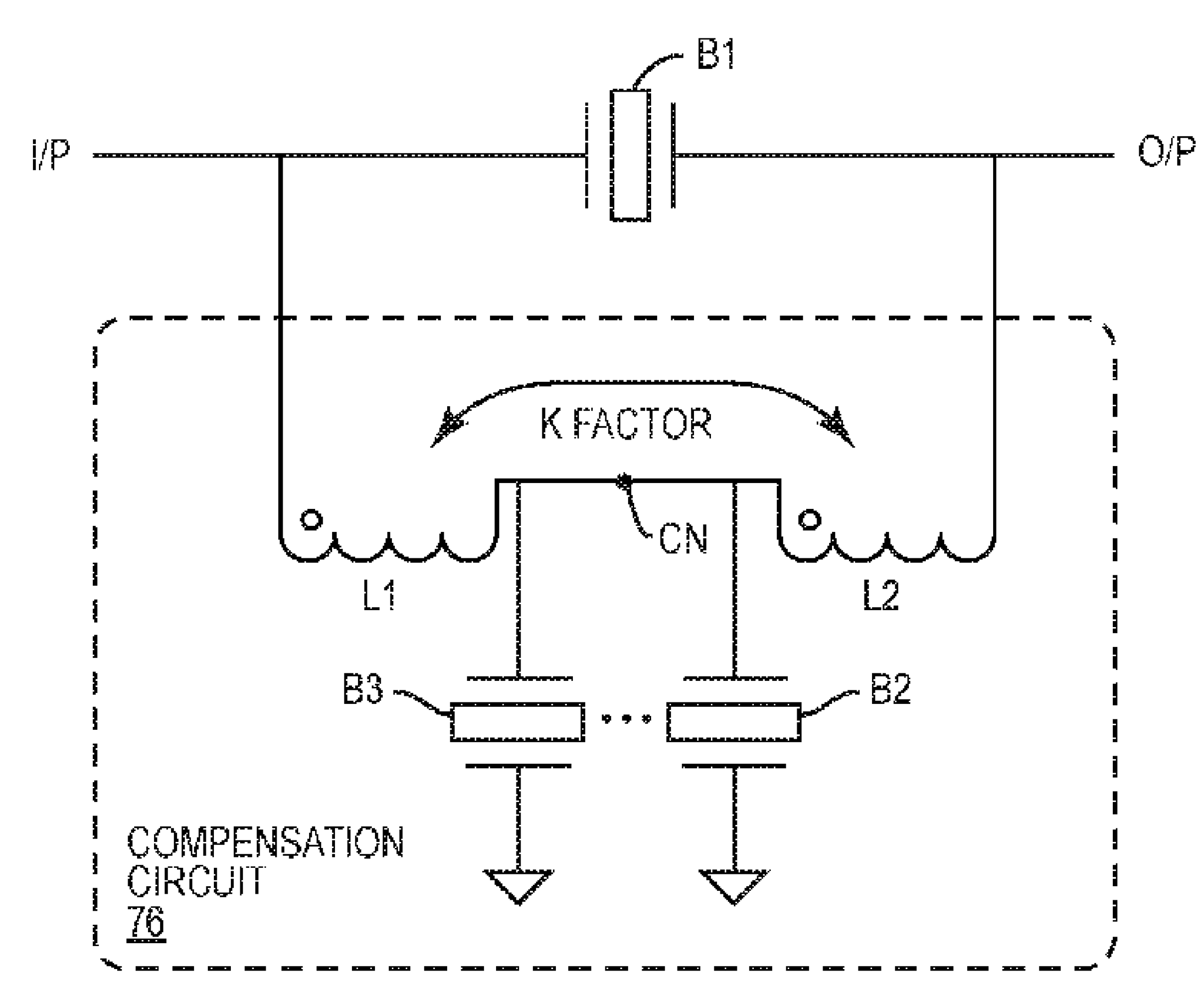
(57) **ABSTRACT**

Filter circuitry is used in communication systems that employ multiple antennas. In general, a communication system may have a transmit path and a receive path. The transmit path extends to a first antenna port and is configured to present signals for transmission in a first communication band and a second communication band to the first antenna port for transmission via a first antenna that is coupled to the first antenna port. The receive path extends to a second antenna port and comprises a first multiple passband/multiple stopband filter that provides a plurality of passbands and a plurality of stopbands interleaved with one another, wherein a first stopband and a second stopband of the plurality of stopbands correspond respectively to the first communication band and the second communication band and are separated by a first passband of the plurality of passbands.

23 Claims, 28 Drawing Sheets



(51) **Int. Cl.**

H04B 1/00	(2006.01)
H04B 11/00	(2006.01)
H03H 9/54	(2006.01)
H03H 9/70	(2006.01)
H03H 9/72	(2006.01)
H01Q 1/24	(2006.01)
H03H 7/09	(2006.01)

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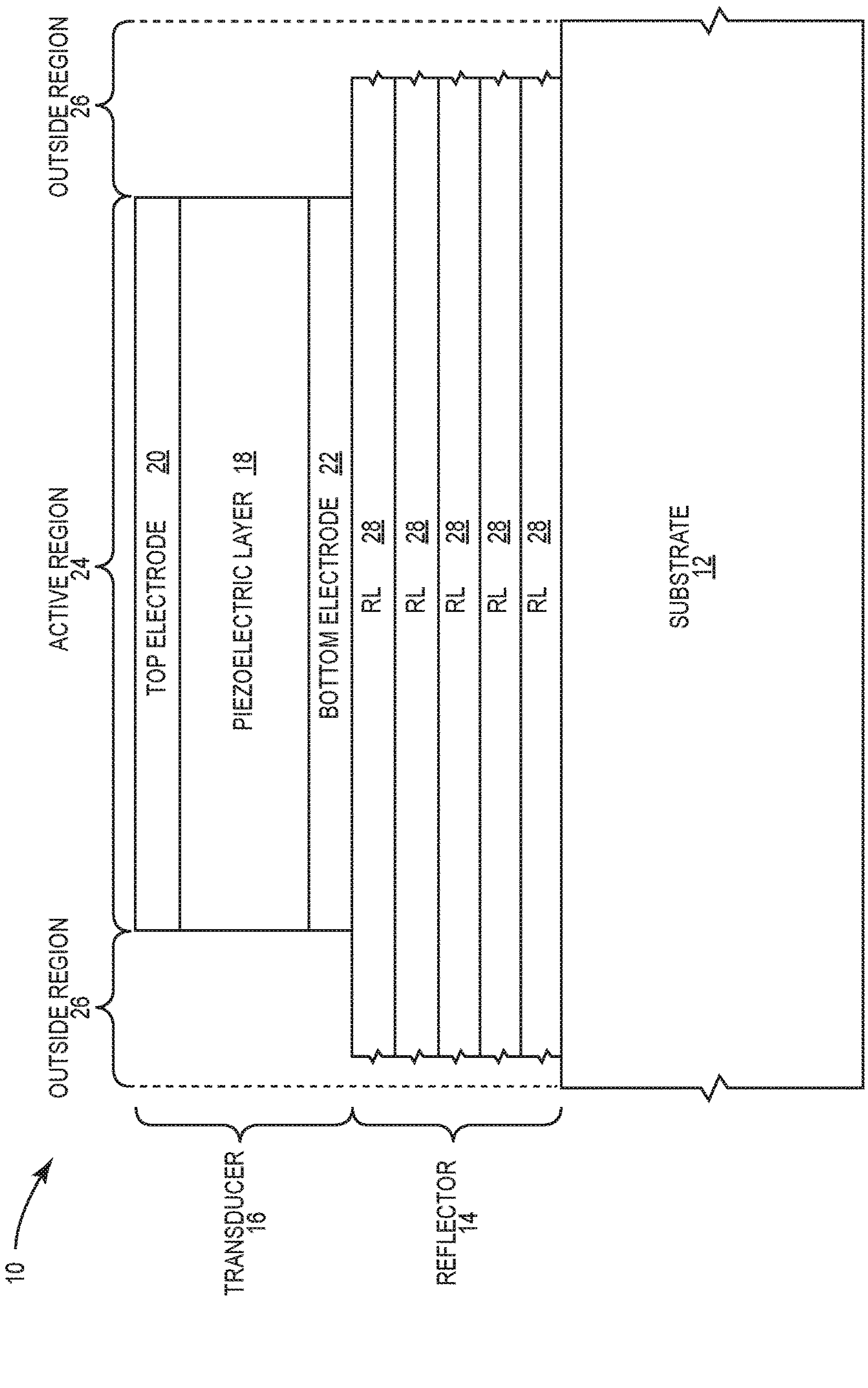


FIG. 1
RELATED ART

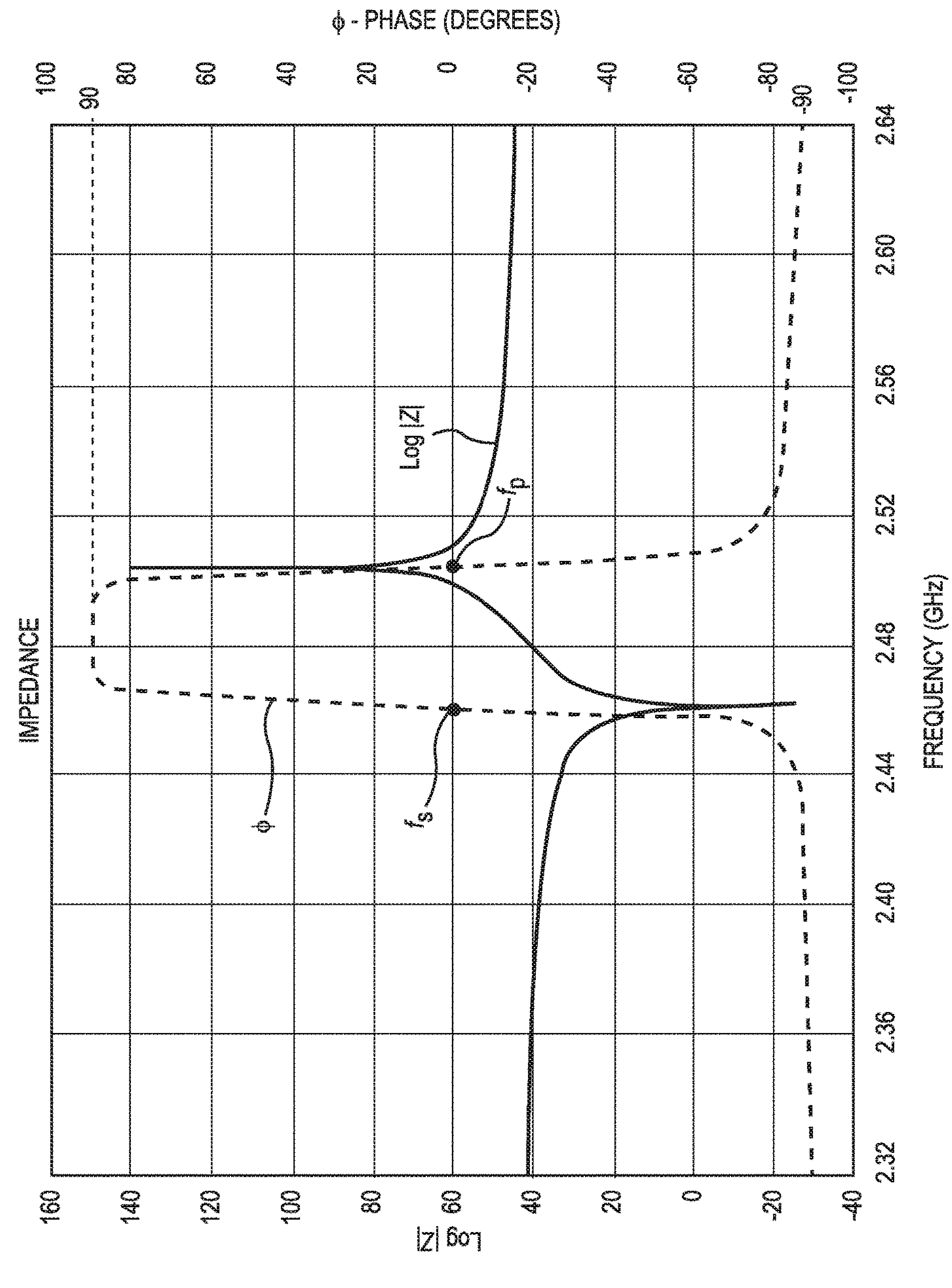

IMPEDANCE
ϕ - PHASE (DEGREES)
100
80
60
40
20
0
-20
-40
-60
-80
-100
90
-90
160
140
120
100
80
60
40
20
0
-20
-40
Log |Z|
FREQUENCY (GHz)
2.32
2.36
2.40
2.44
2.48
2.52
2.56
2.60
2.64
Log |Z|
f_p
f_s
ϕ

FIG. 2

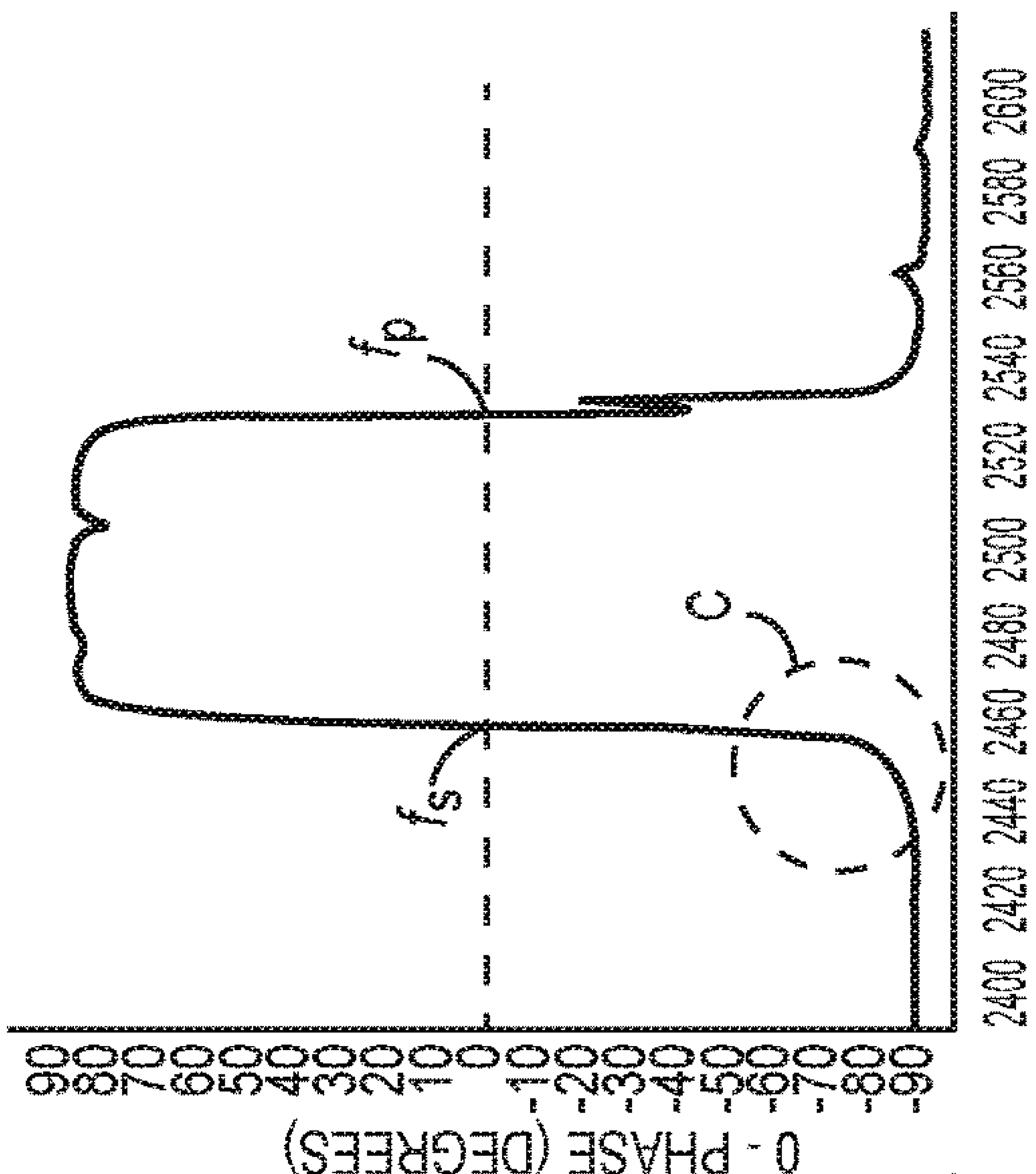


FIG. 3C

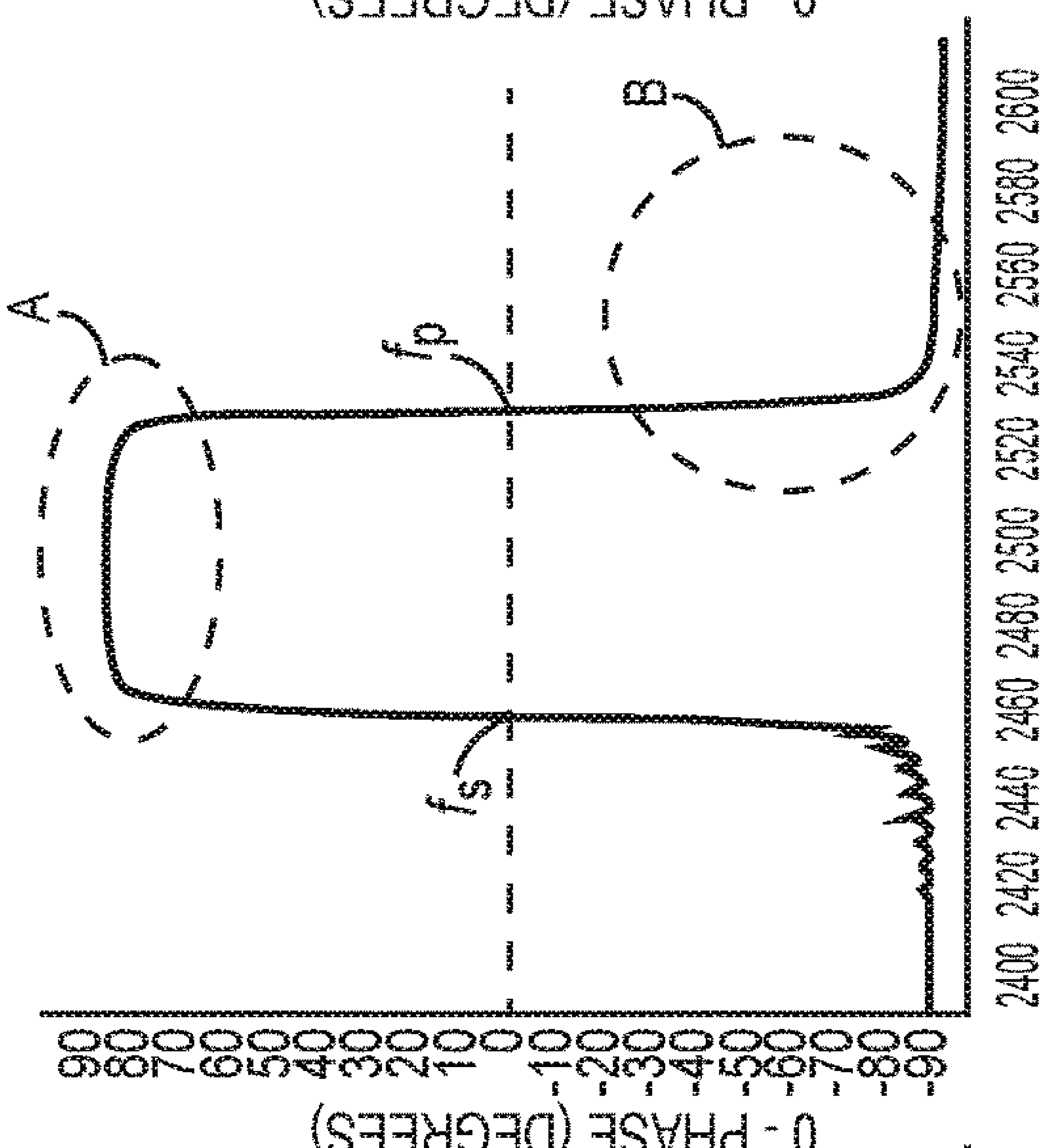


FIG. 3B

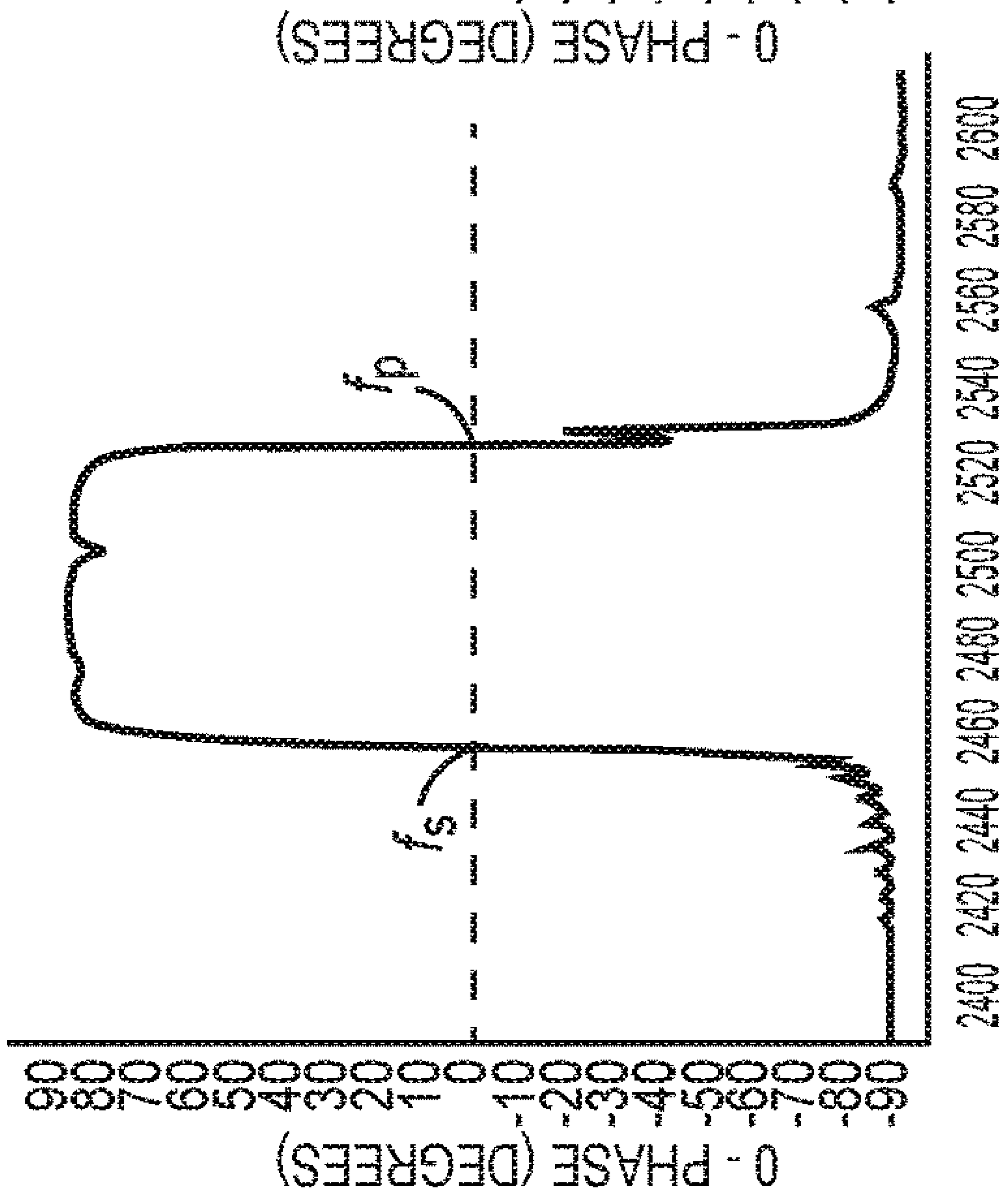


FIG. 3A

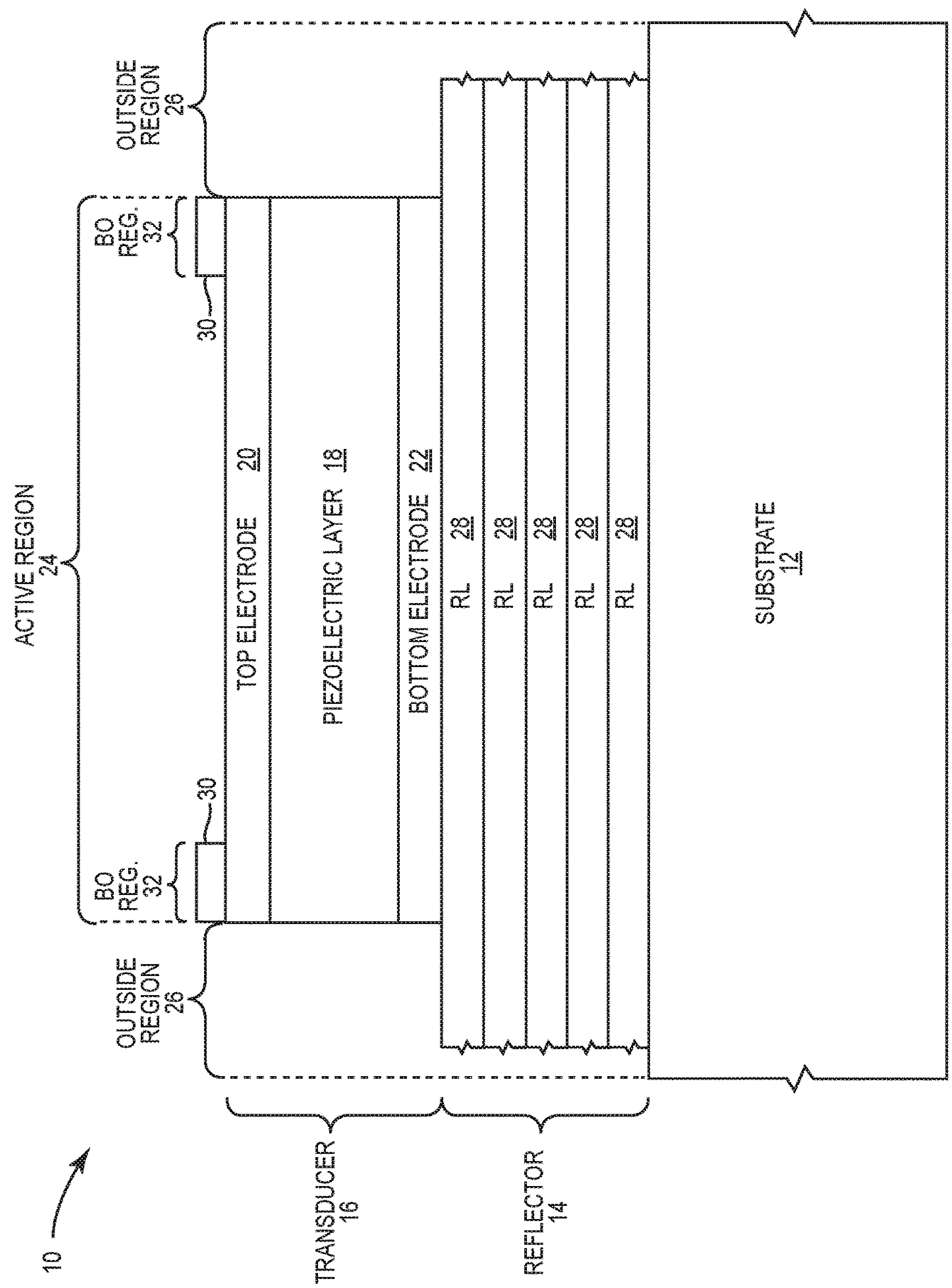


FIG. 4
RELATED ART

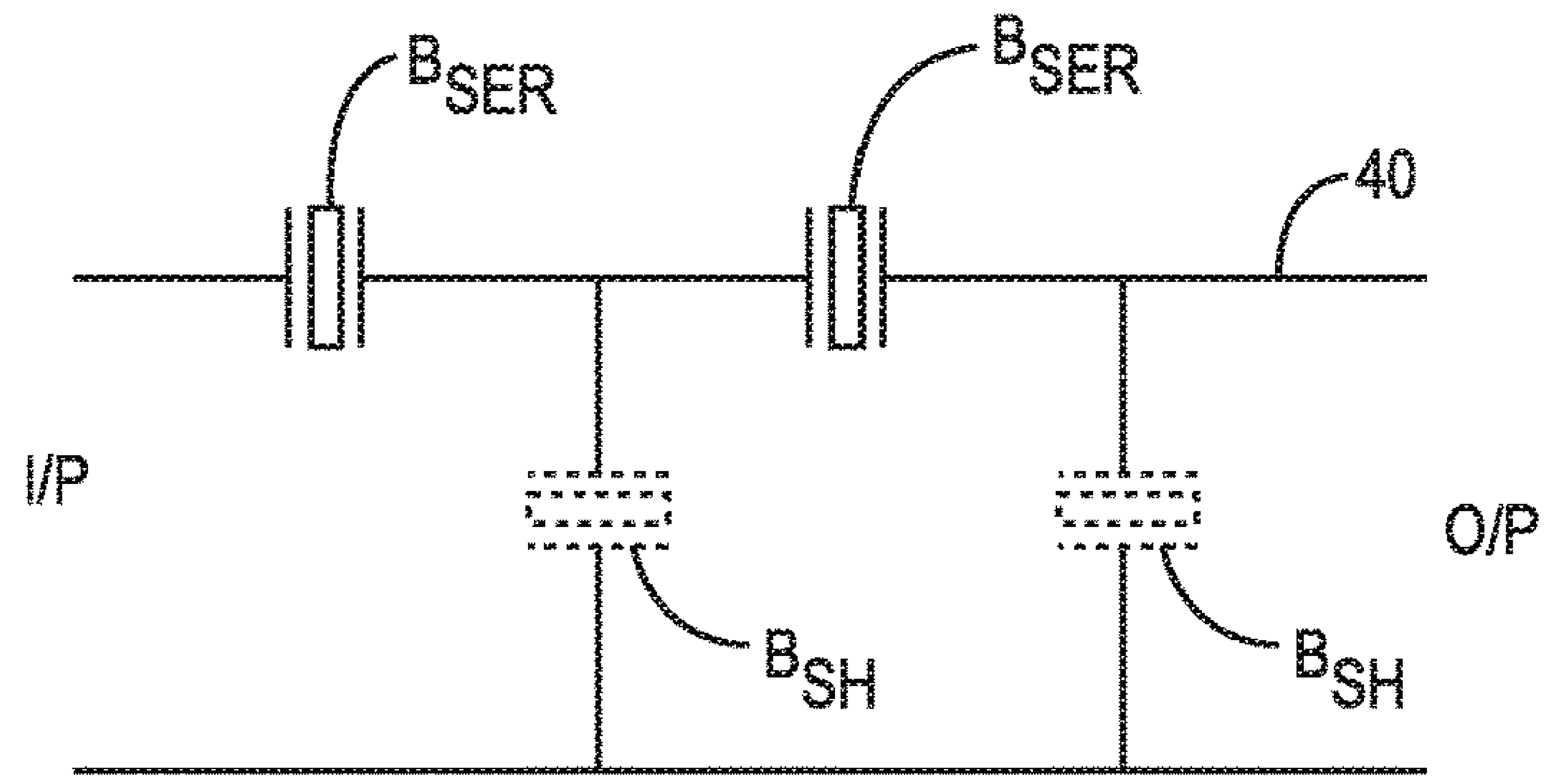


FIG. 5A

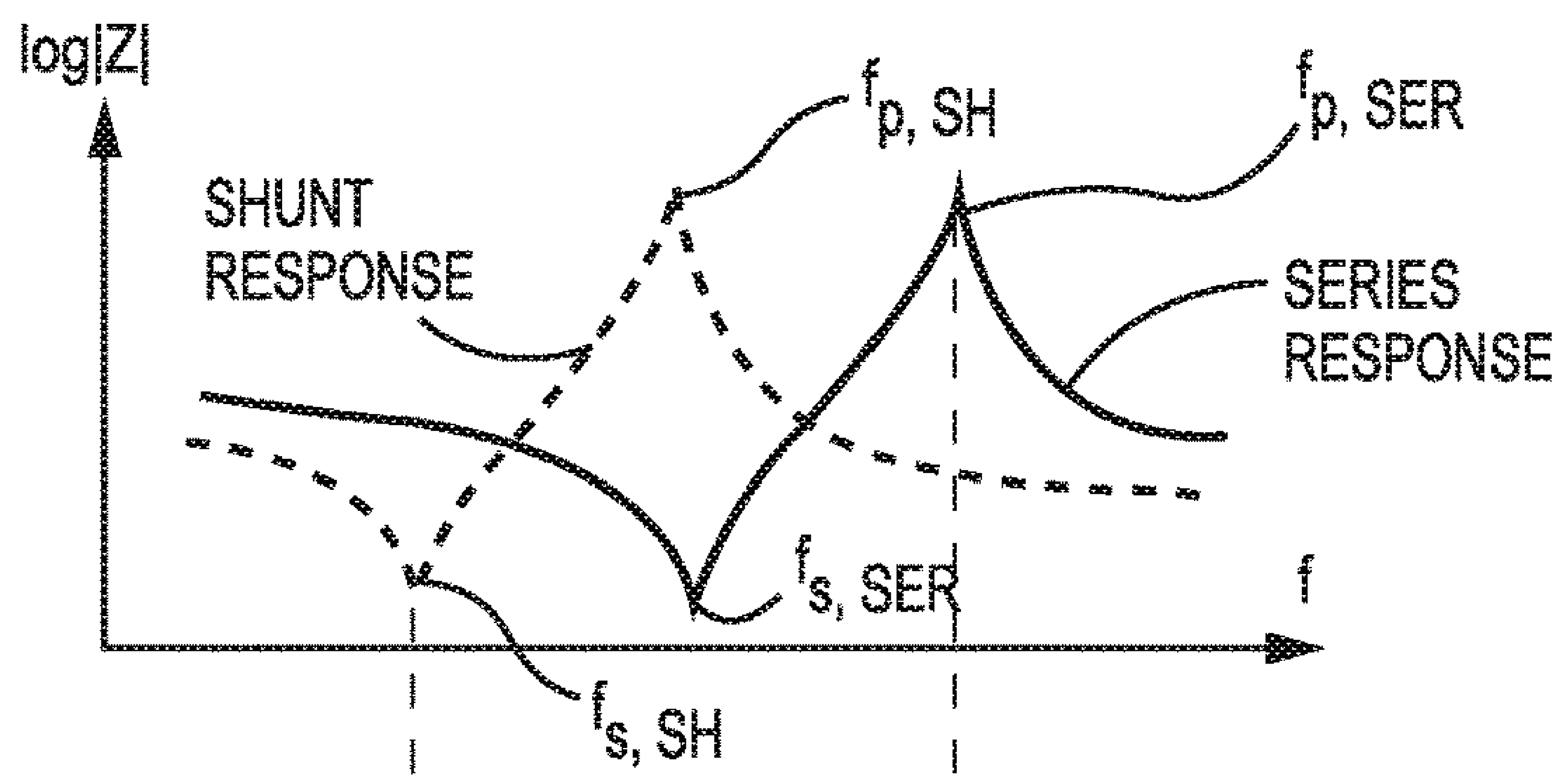


FIG. 5B

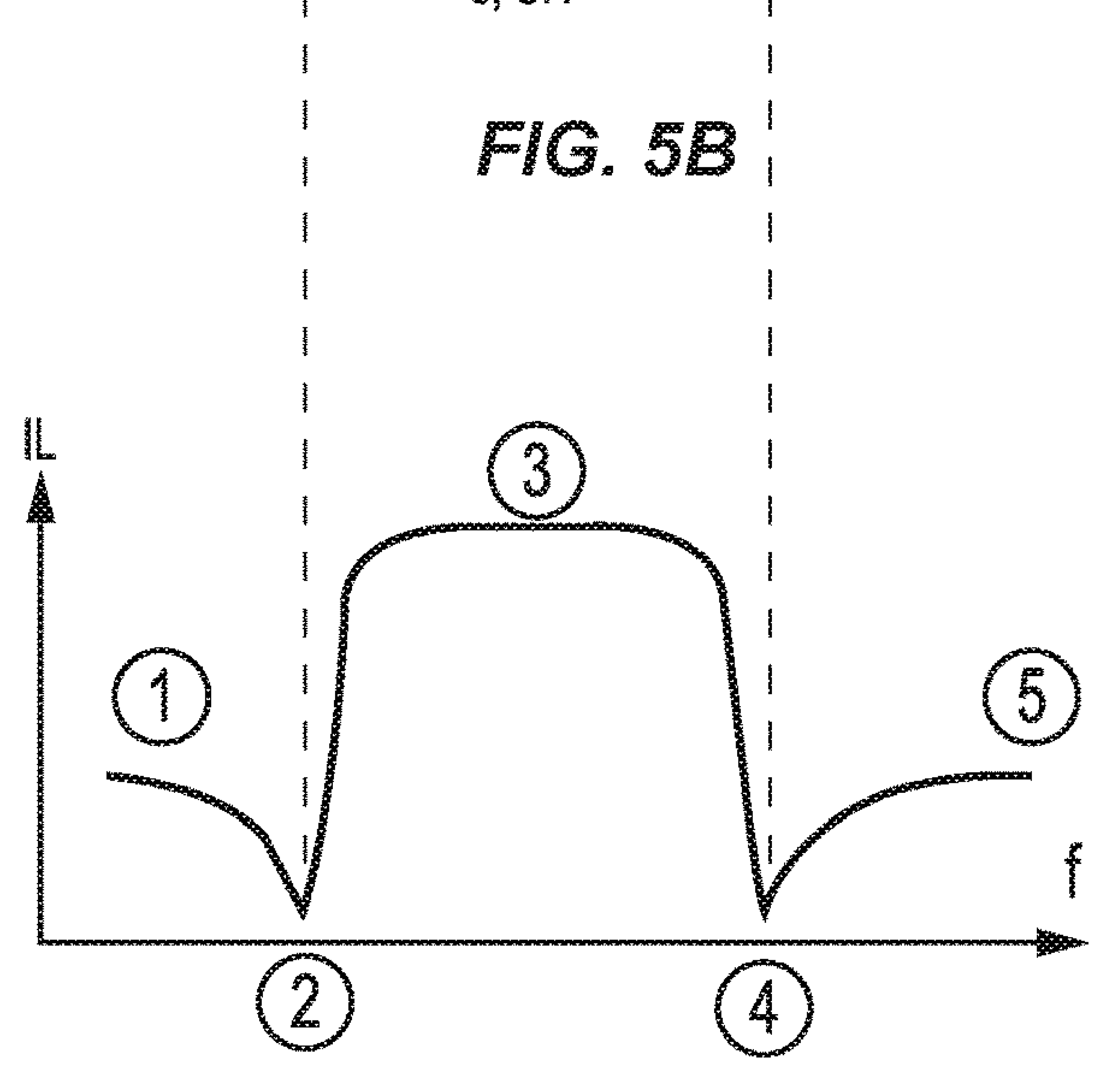


FIG. 5C

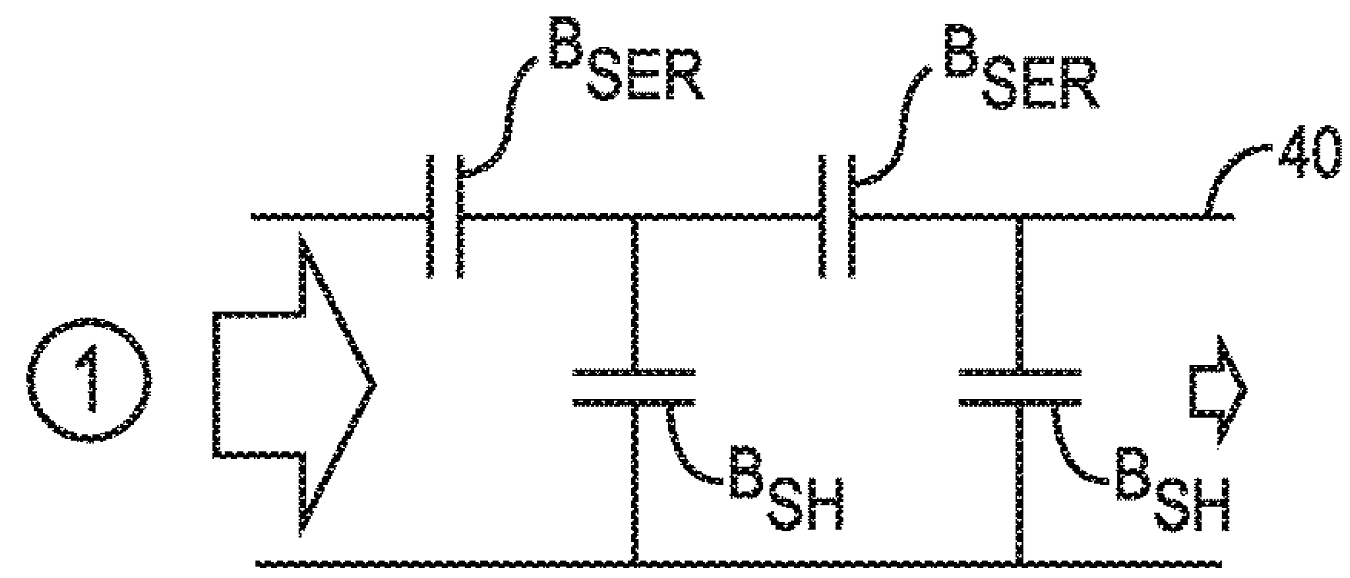


FIG. 6A

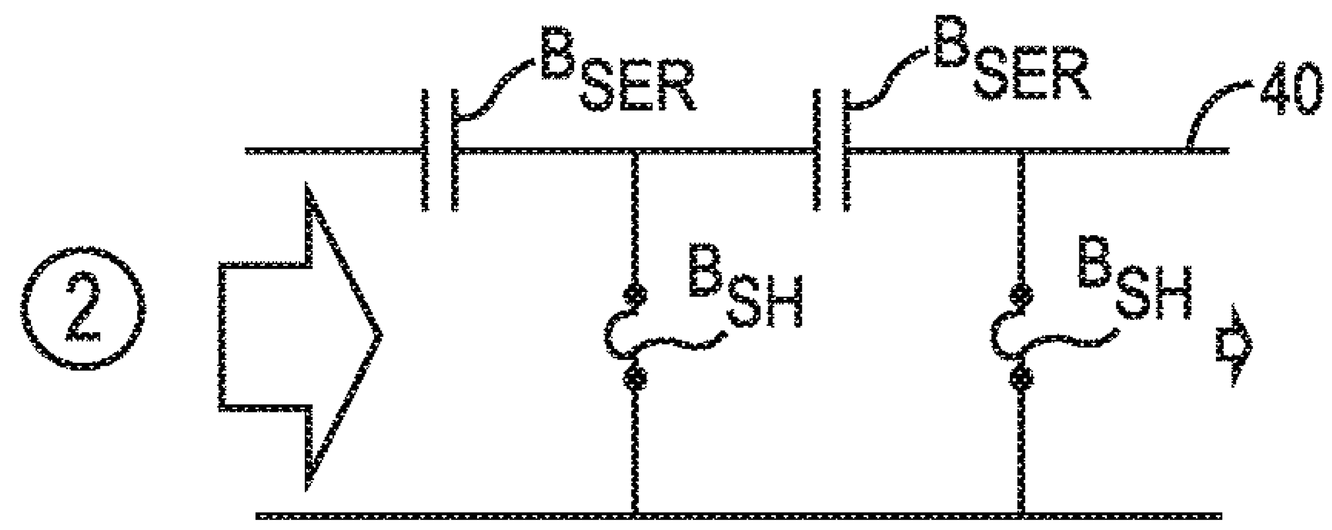


FIG. 6B

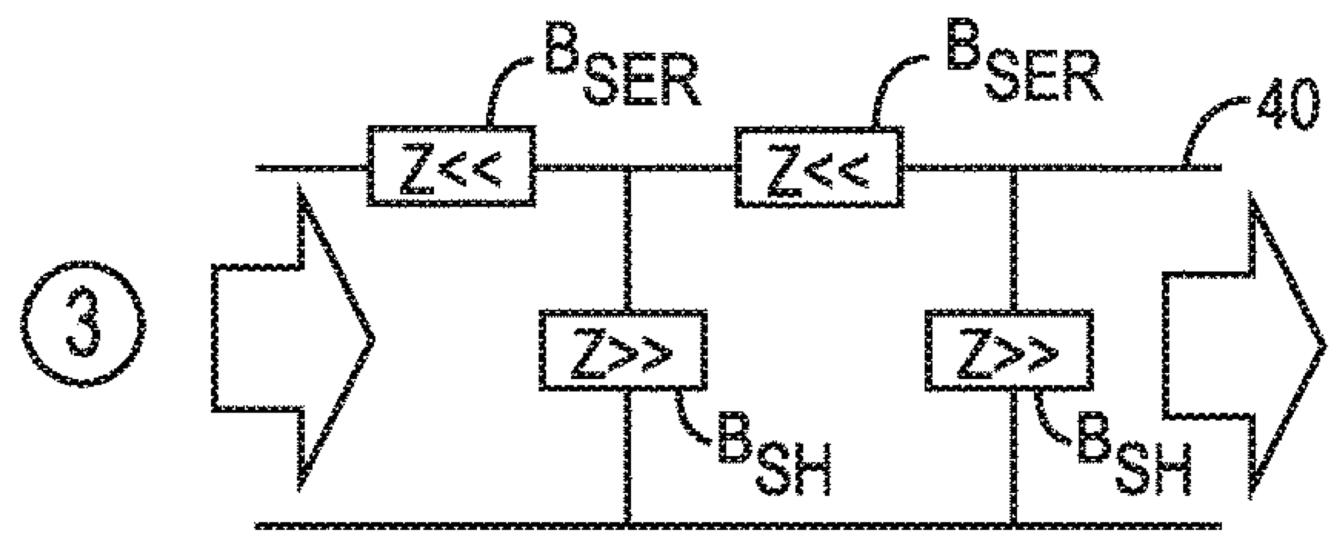


FIG. 6C

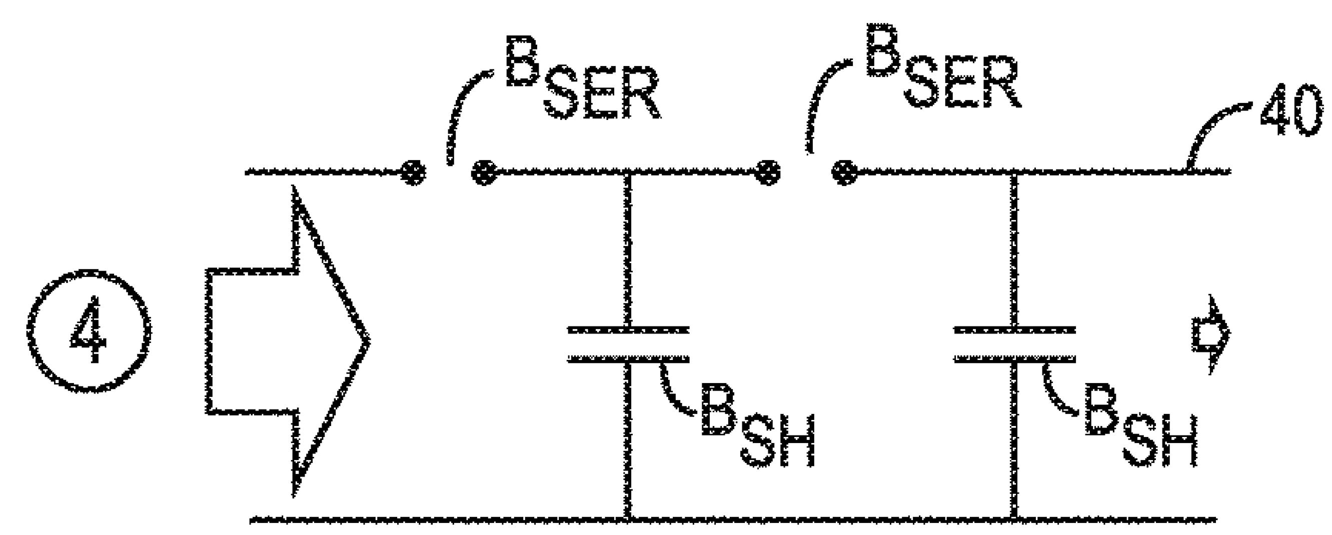


FIG. 6D

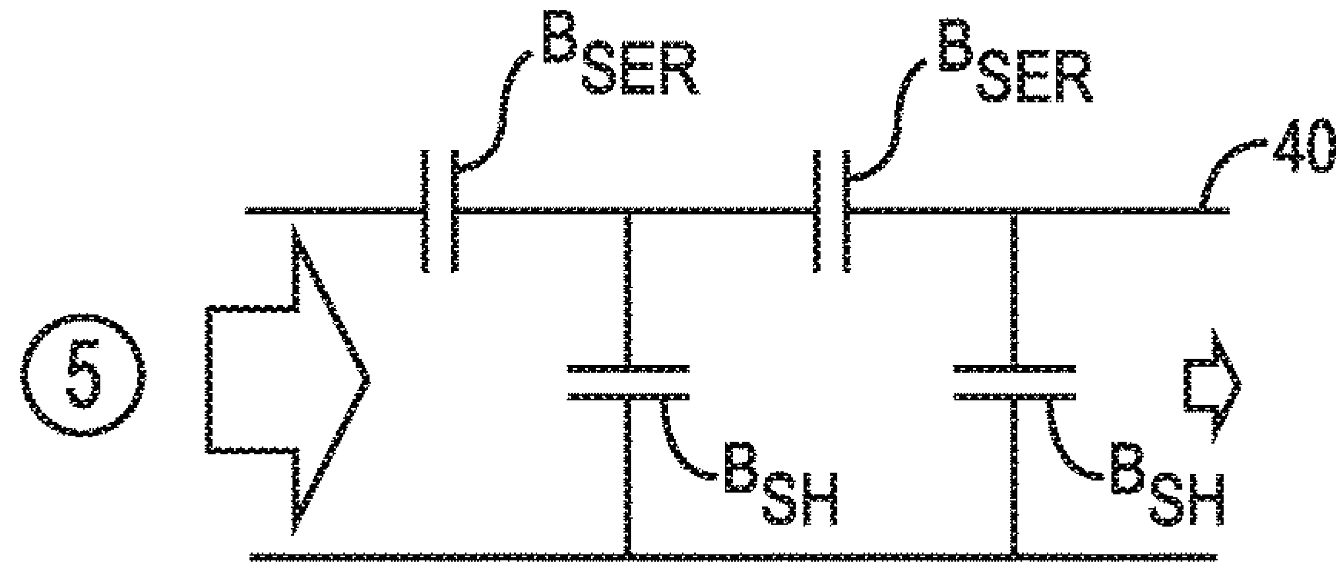


FIG. 6E

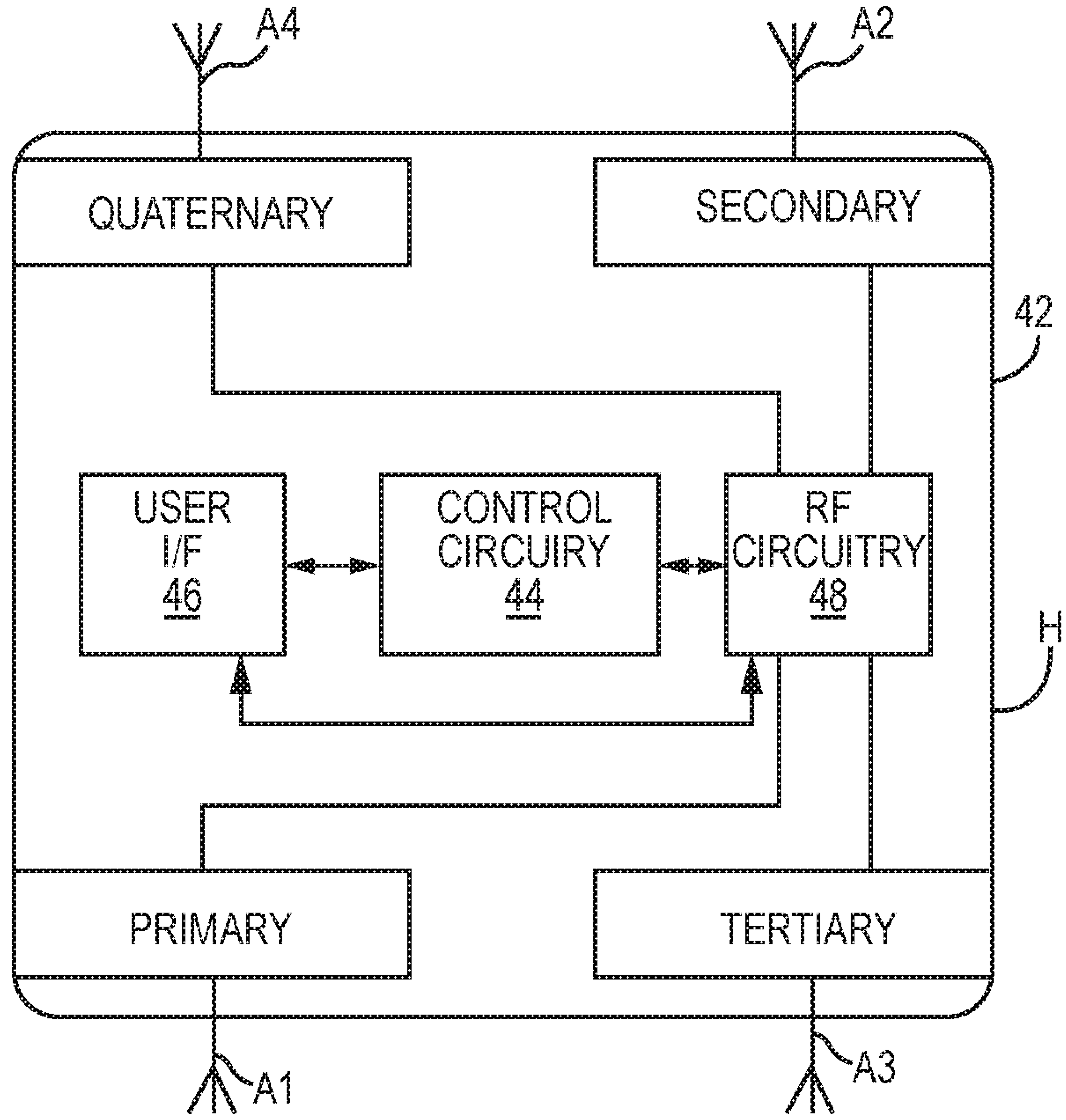

A4
A2
QUATERNARY
SECONDARY
42
USER
I/F
46
CONTROL
CIRCUIRY
44
RF
CIRCUITRY
48
H
PRIMARY
TERTIARY
A1
A3

FIG. 7

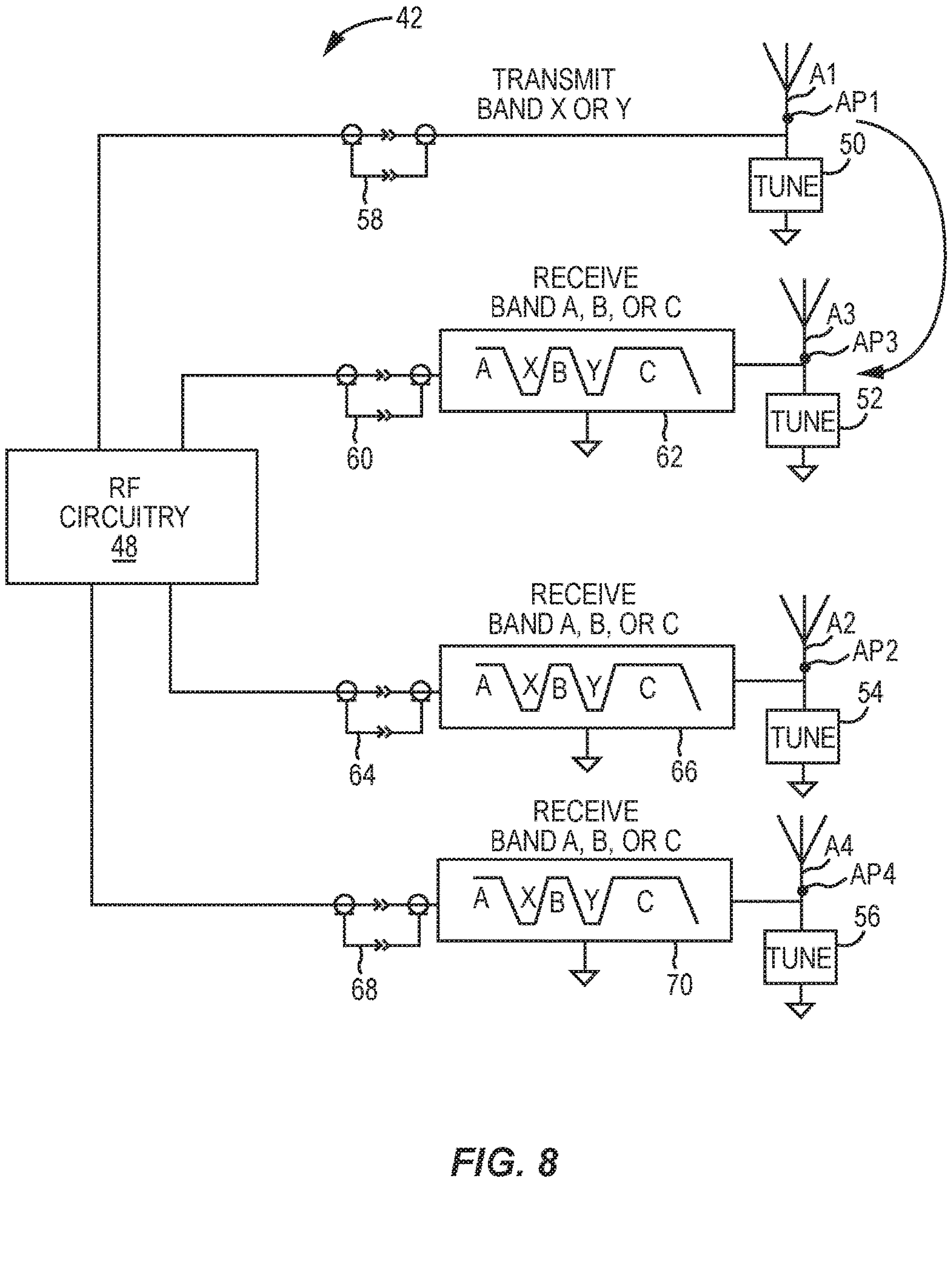

42
TRANSMIT
BAND X OR Y
A1
AP1
50
TUNE
58
RECEIVE
BAND A, B, OR C
A3
AP3
A X B Y C
TUNE
52
60
62
RF
CIRCUITRY
48
RECEIVE
BAND A, B, OR C
A2
AP2
A X B Y C
54
TUNE
64
66
RECEIVE
BAND A, B, OR C
A4
AP4
A X B Y C
56
TUNE
68
70

FIG. 8

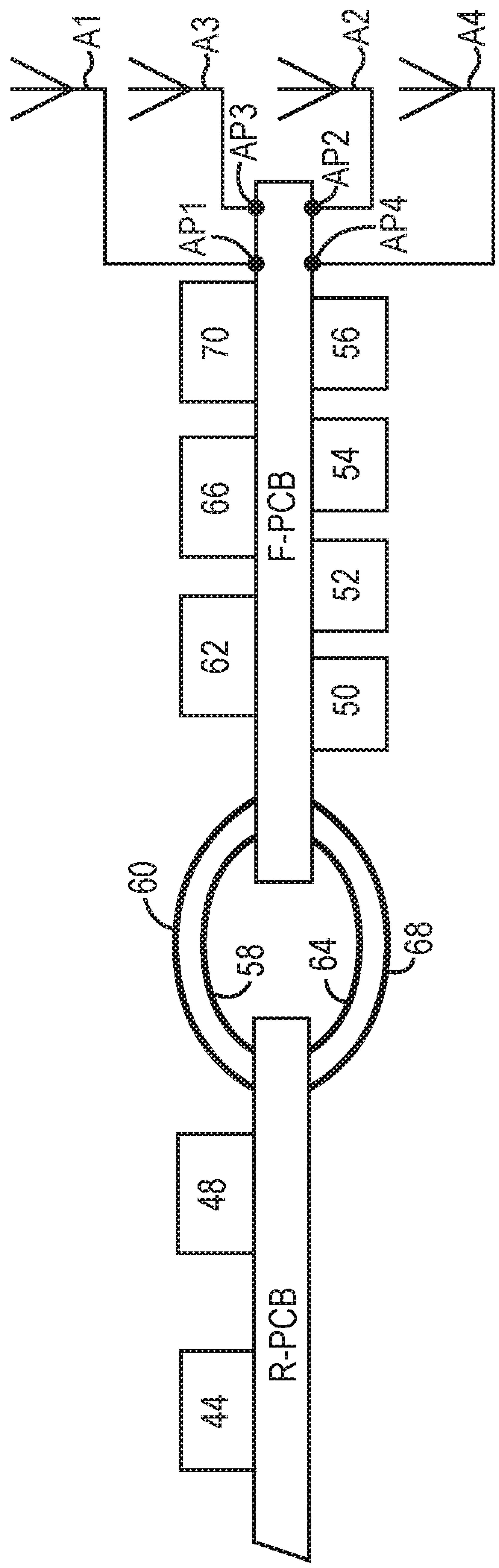
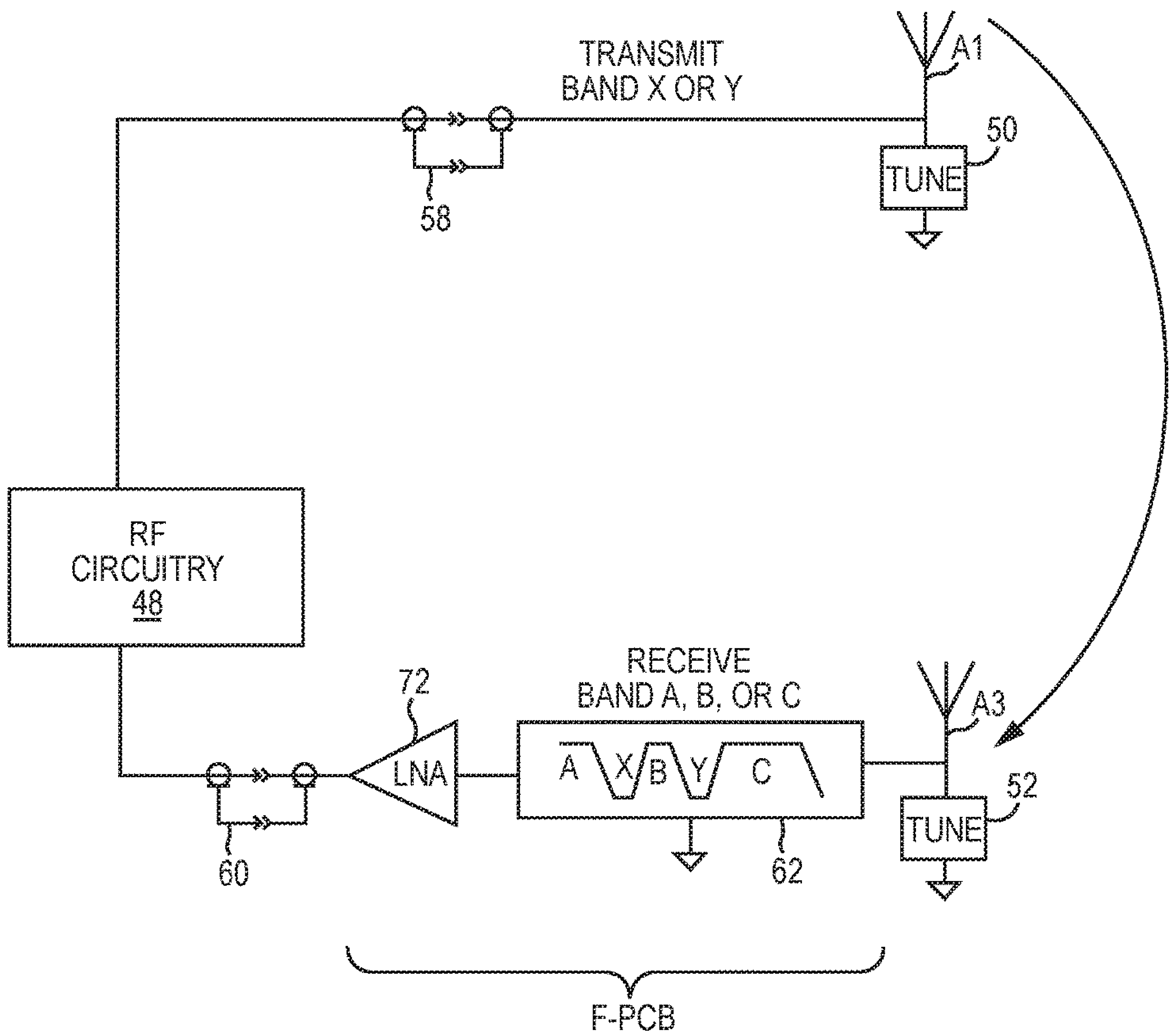

A1
A3
A2
A4
AP3
AP2
AP1
AP4
70
66
62
56
54
52
50
F-PCB
60
58
64
68
48
44
R-PCB

FIG. 9



TRANSMIT
BAND X OR Y
A1
50
TUNE
58
RF
CIRCUITRY
48
72
RECEIVE
BAND A, B, OR C
A3
LNA
A
X
B
Y
C
52
TUNE
60
62
F-PCB

FIG. 10

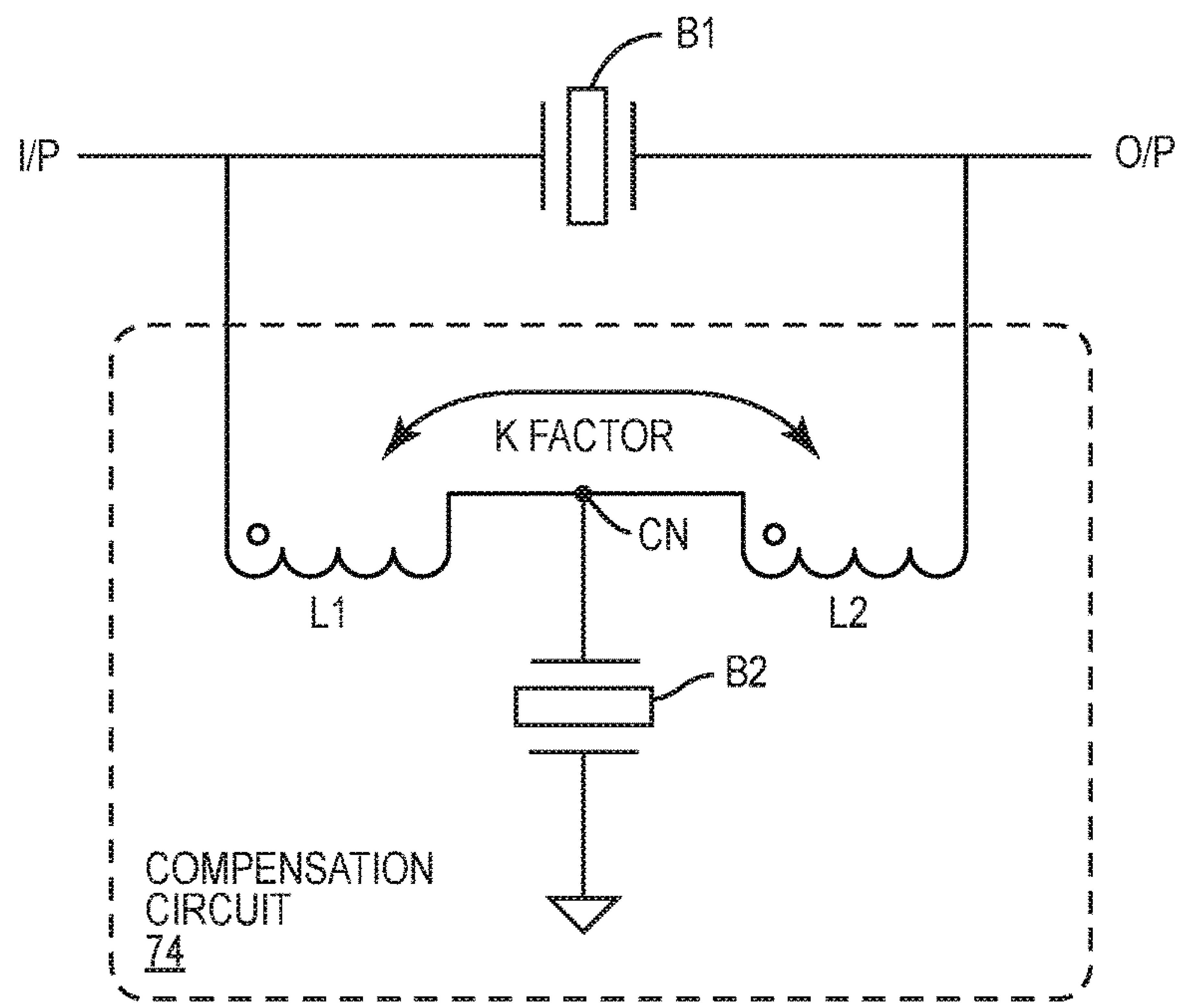

B1
I/P
O/P
K FACTOR
CN
L1
L2
B2
COMPENSATION
CIRCUIT
74

FIG. 11

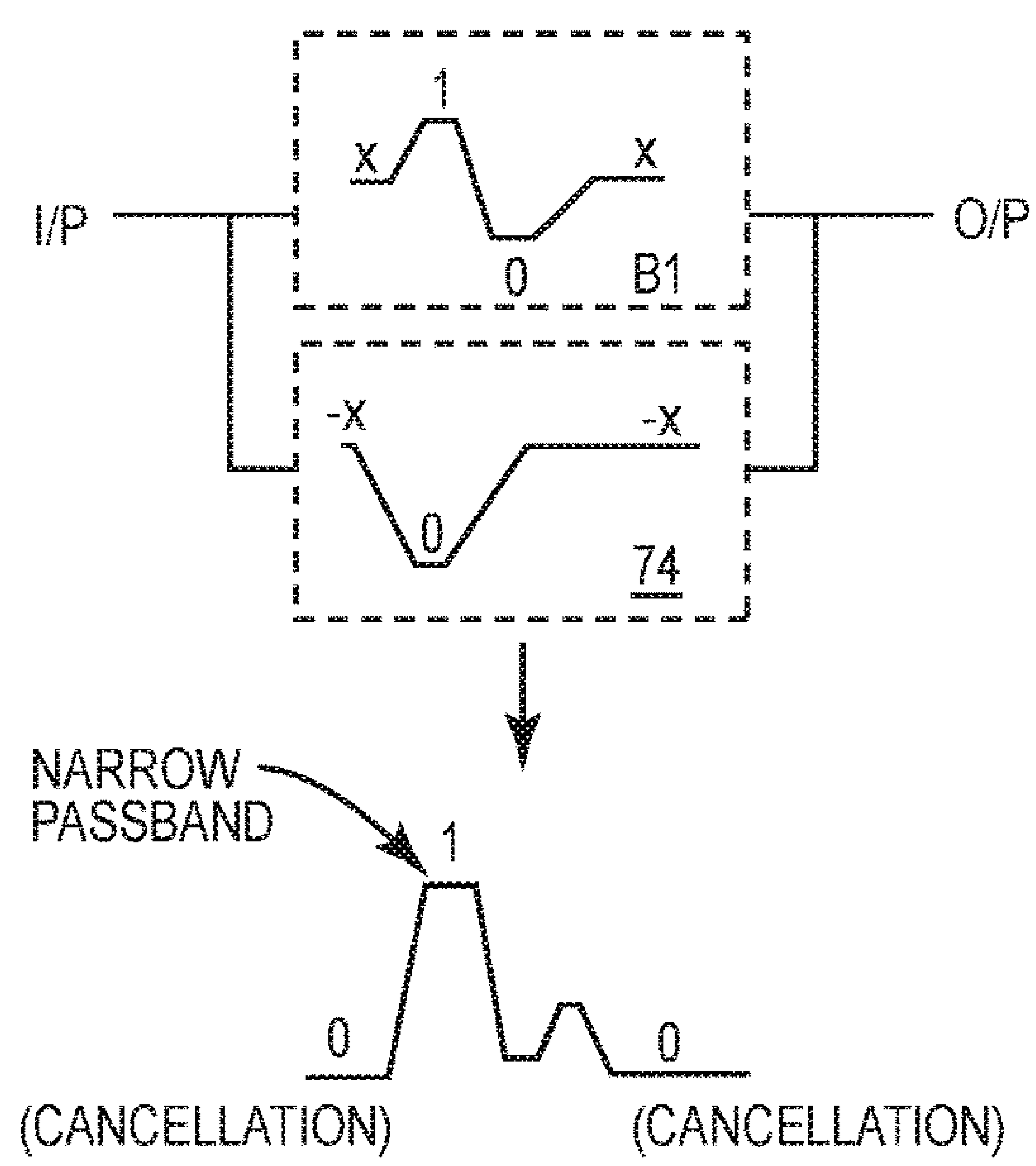

I/P
O/P
x
1
x
0
B1
-x
-x
0
74
NARROW
PASSBAND
1
0
0
(CANCELLATION)
(CANCELLATION)

FIG. 12

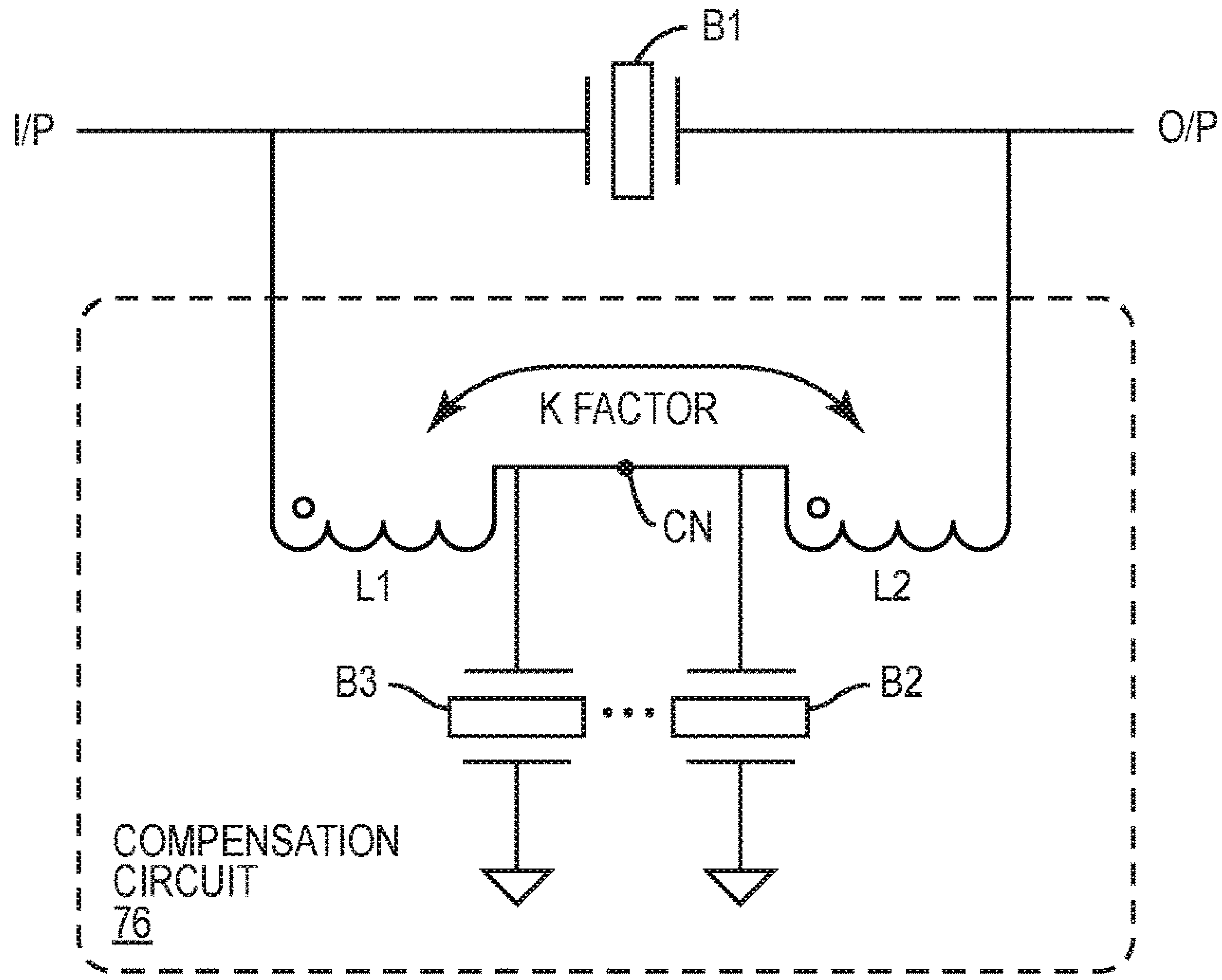

B1
I/P
O/P
K FACTOR
CN
L1
L2
B3
B2
COMPENSATION
CIRCUIT
76

FIG. 13

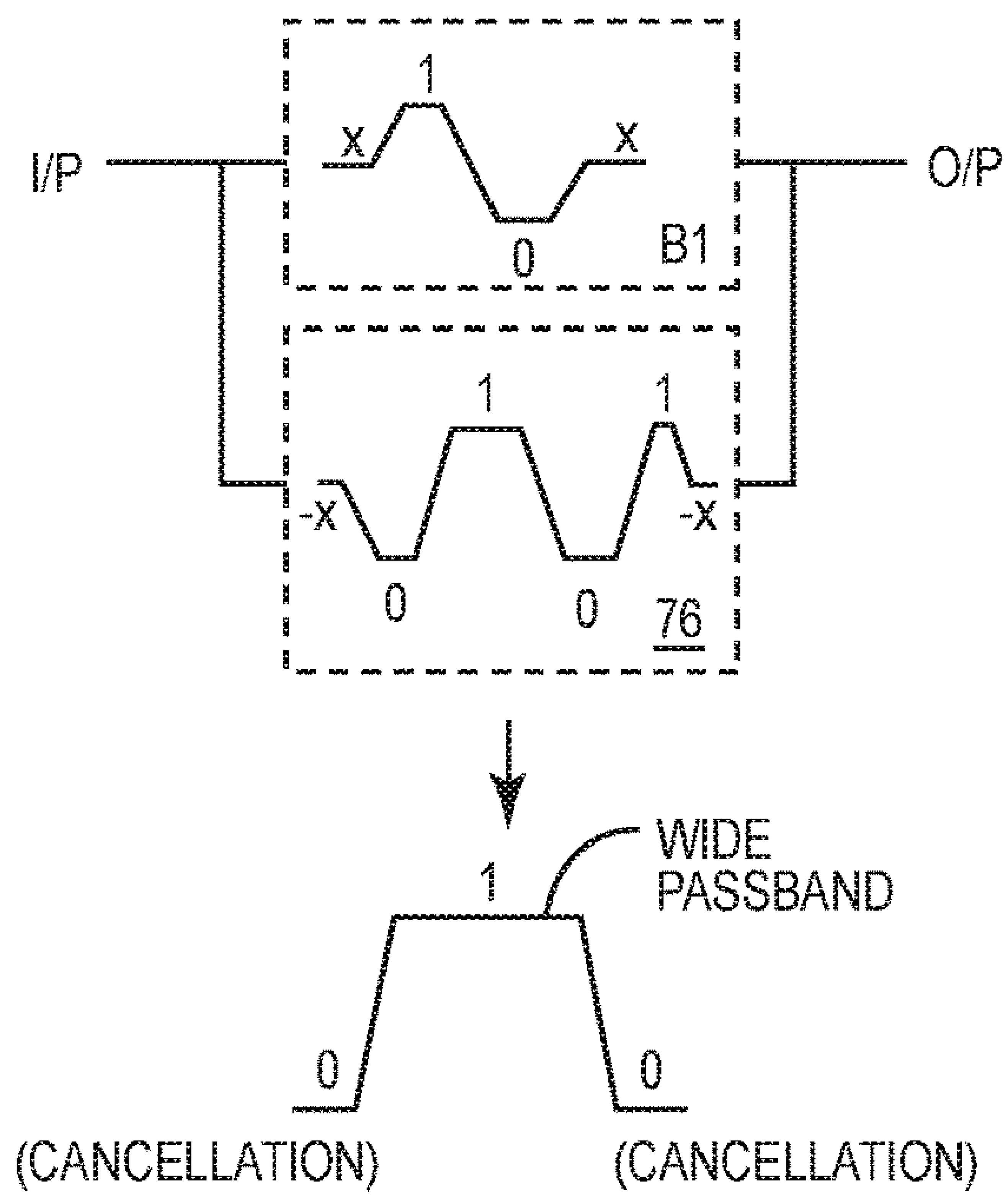

I/P
O/P
1
x
x
0
B1
1
1
-x
-x
0
0
76
WIDE
PASSBAND
1
0
0
(CANCELLATION)
(CANCELLATION)

FIG. 14

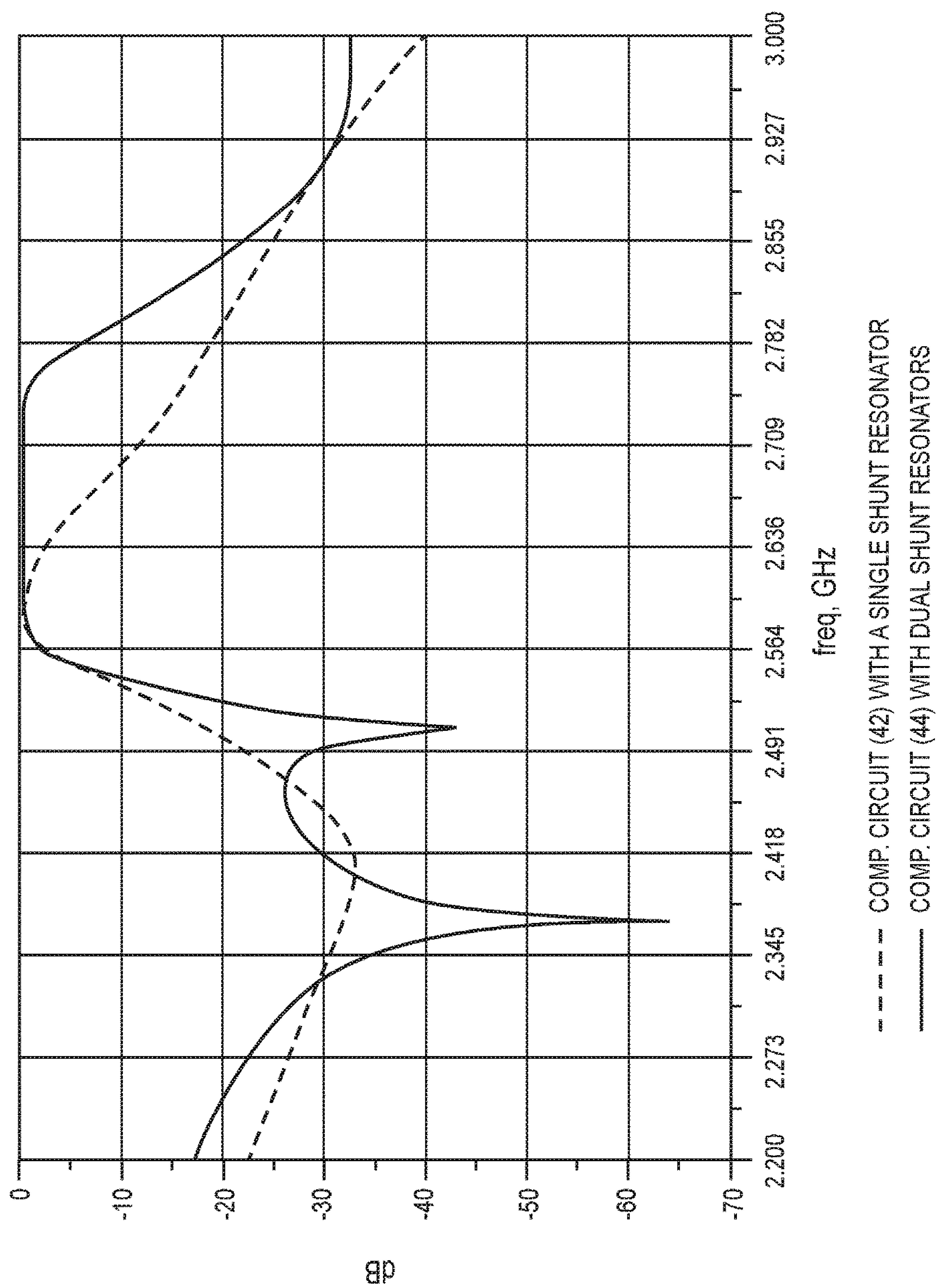

dB
0
-10
-20
-30
-40
-50
-60
-70
2.200
2.273
2.345
2.418
2.491
2.564
2.636
2.709
2.782
2.855
2.927
3.000
freq, GHz
COMP. CIRCUIT (42) WITH A SINGLE SHUNT RESONATOR
COMP. CIRCUIT (44) WITH DUAL SHUNT RESONATORS

FIG. 15

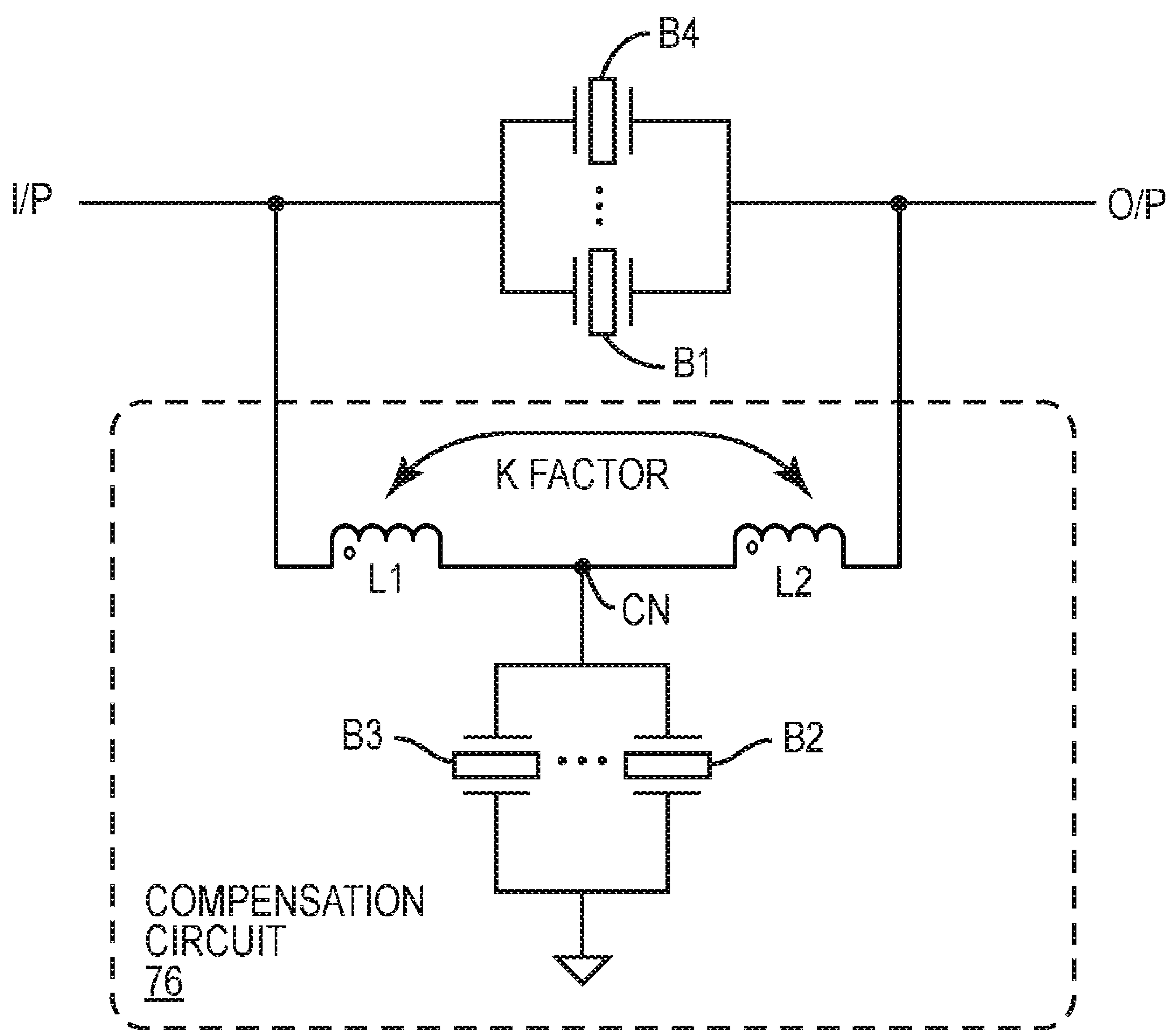


FIG. 16

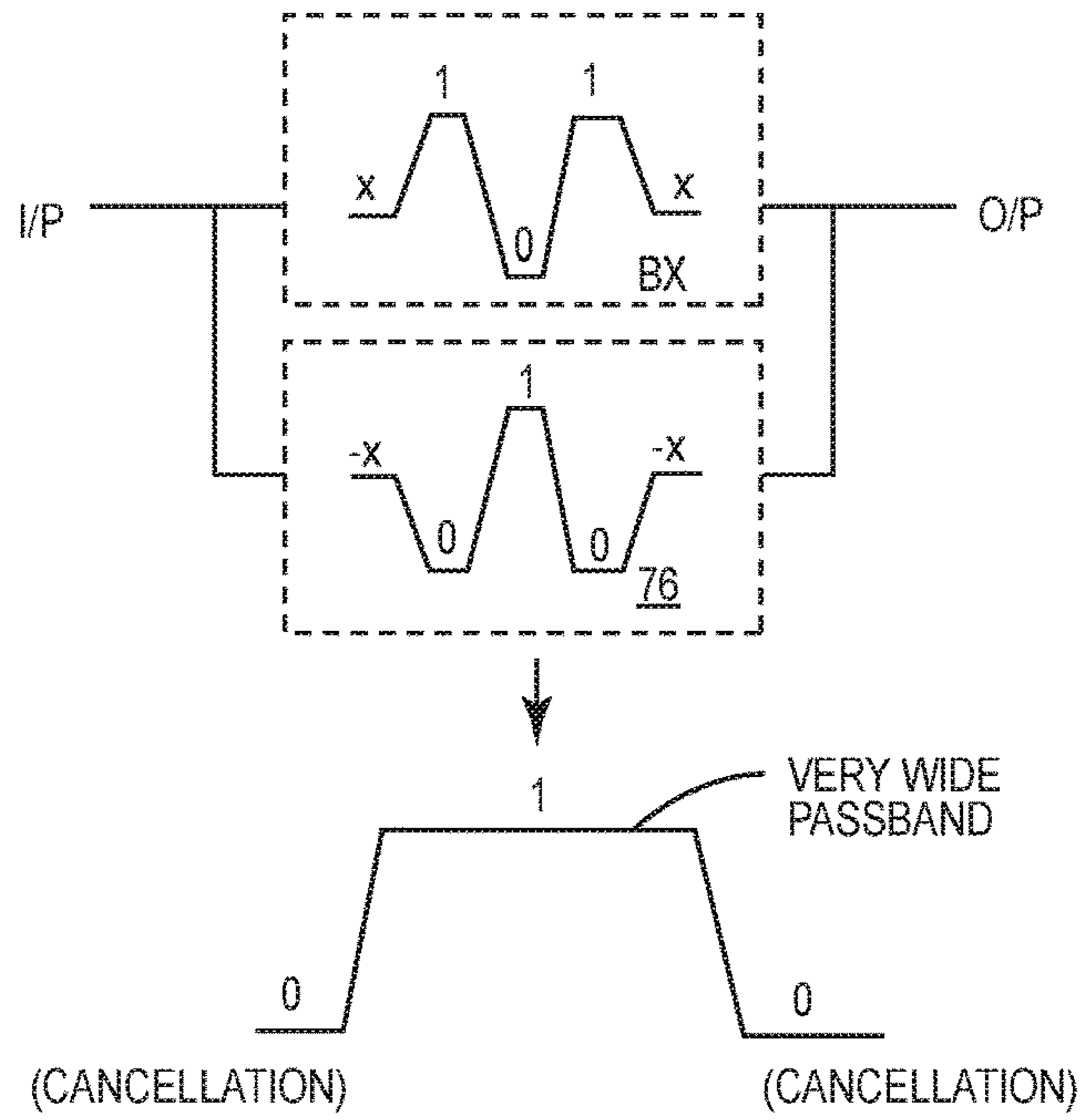

1
1
x
x
I/P
O/P
0
BX
1
-x
-x
0
0
76
1
VERY WIDE PASSBAND
0
0
(CANCELLATION)
(CANCELLATION)

FIG. 17

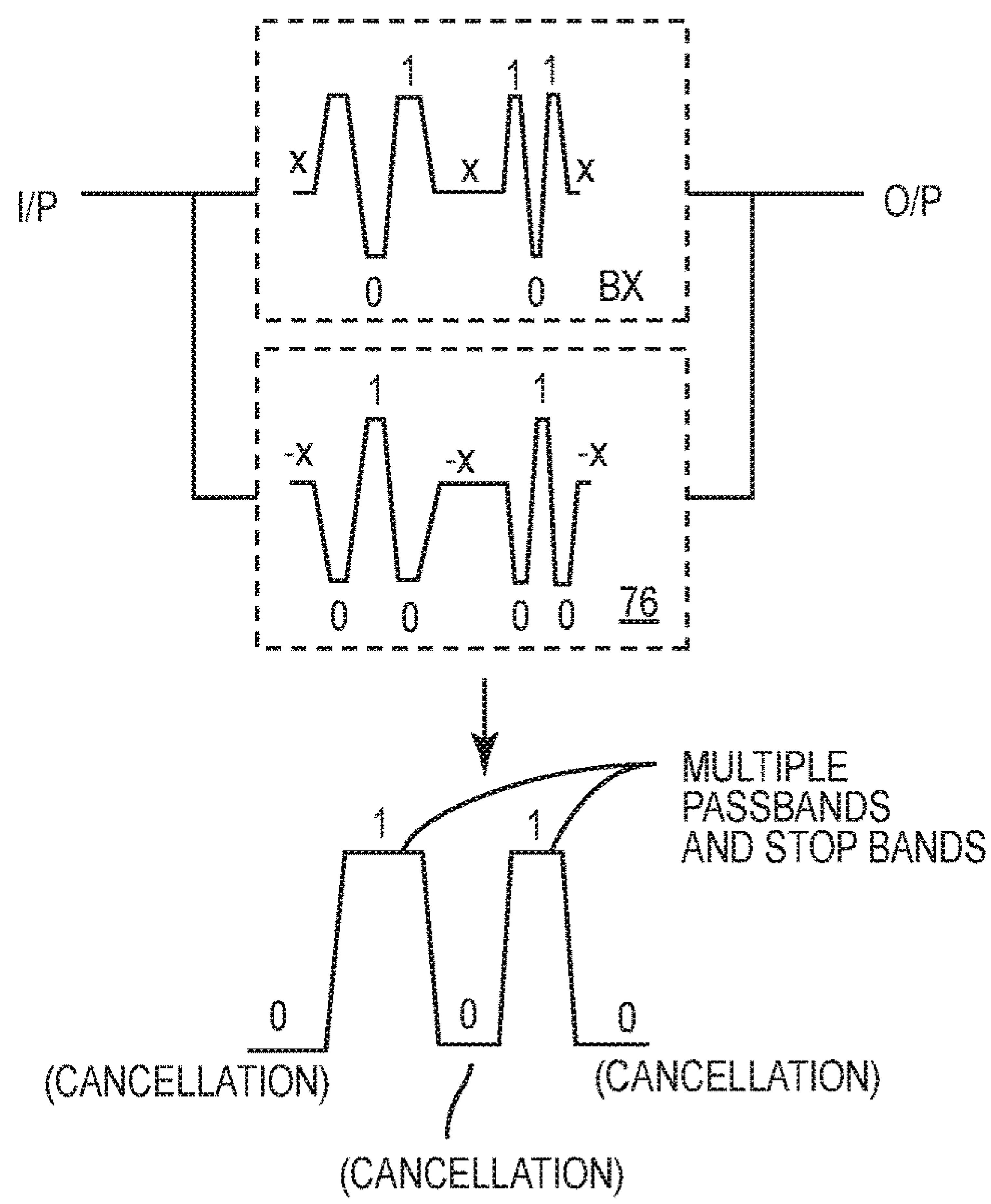

1
1 1
x
x
x
I/P
O/P
0
0
BX
1
1
-x
-x
-x
0 0
0 0
76
MULTIPLE PASSBANDS AND STOP BANDS
1
1
0
0
0
(CANCELLATION)
(CANCELLATION)
(CANCELLATION)

FIG. 18

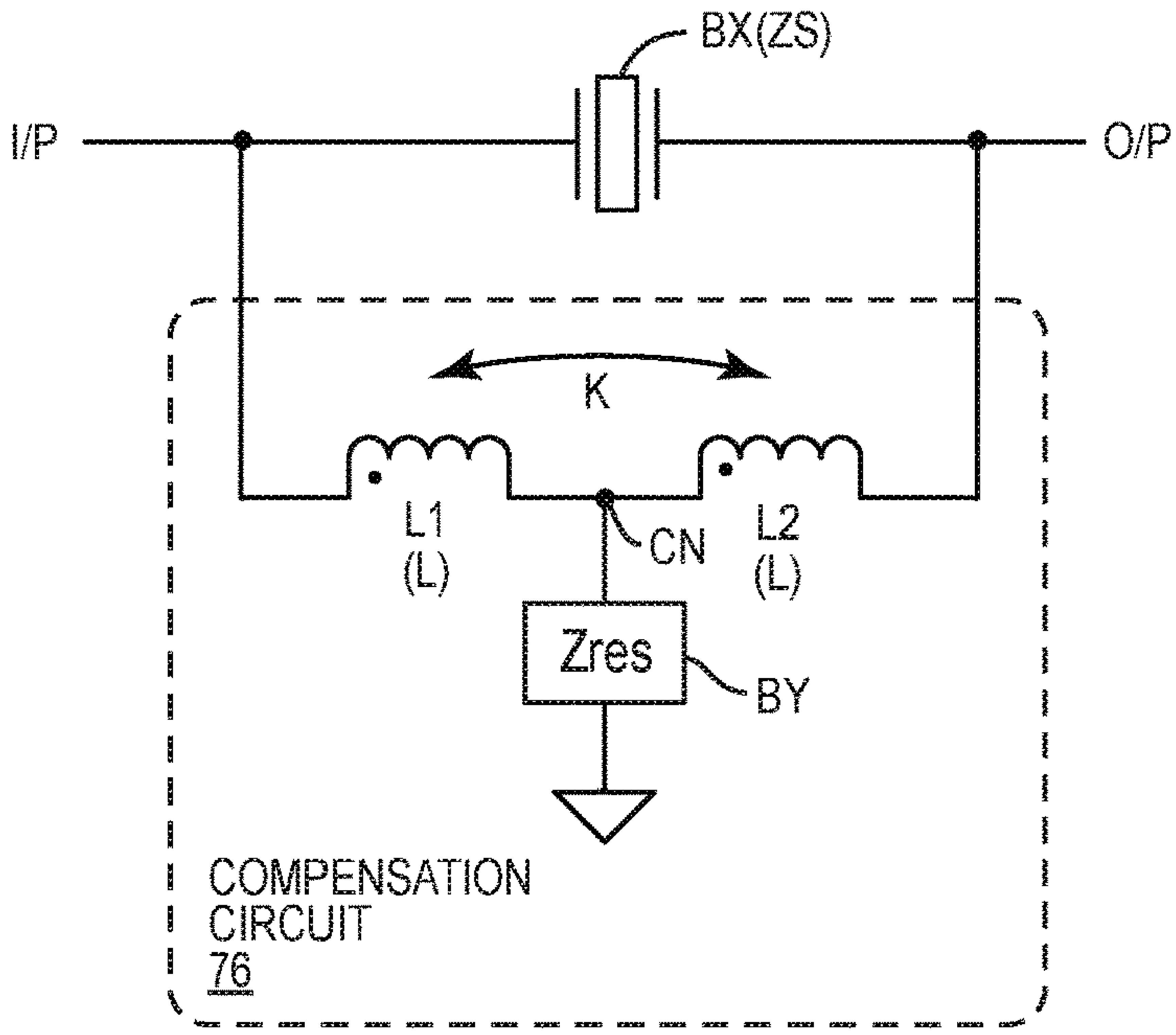


FIG. 19A

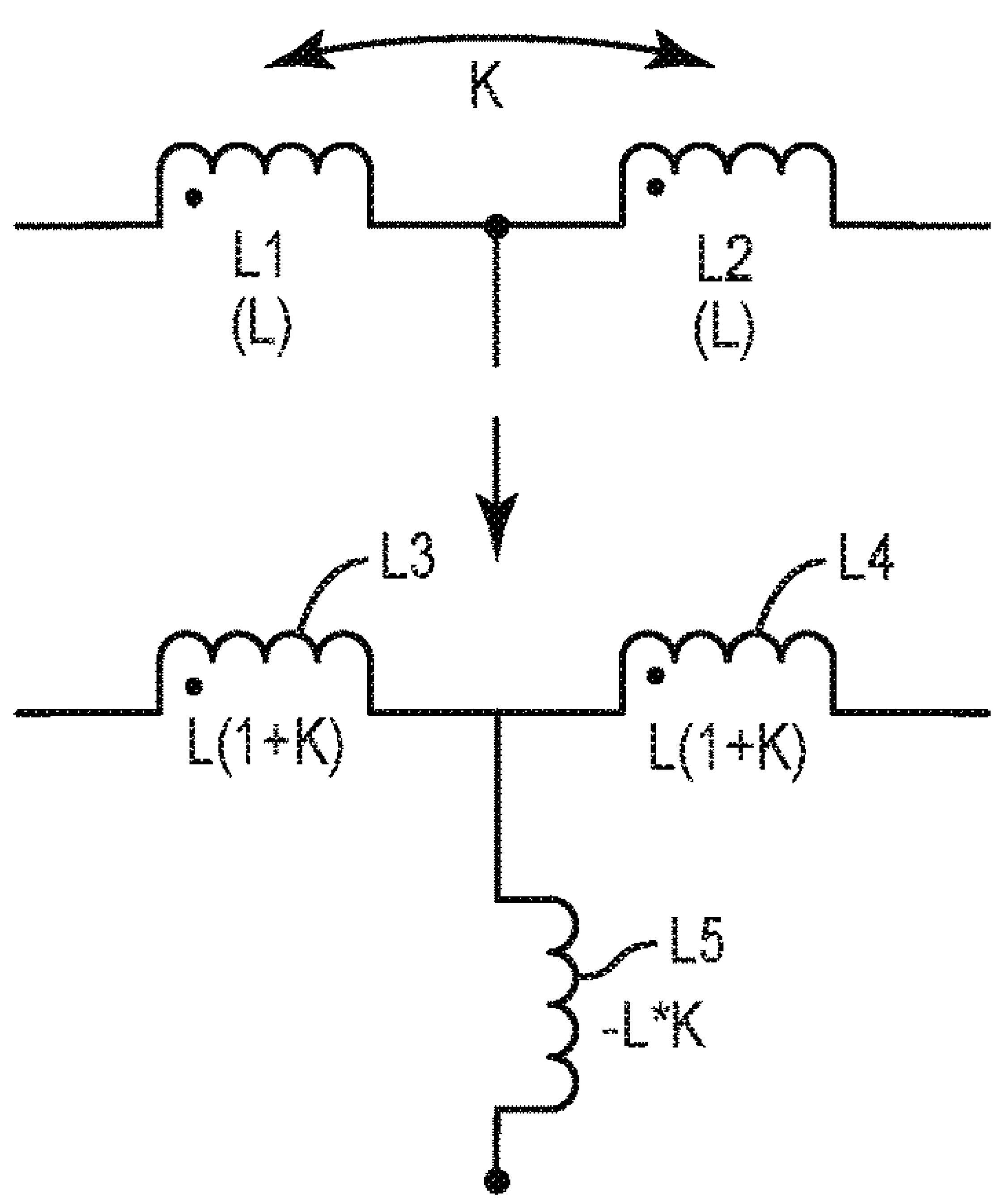
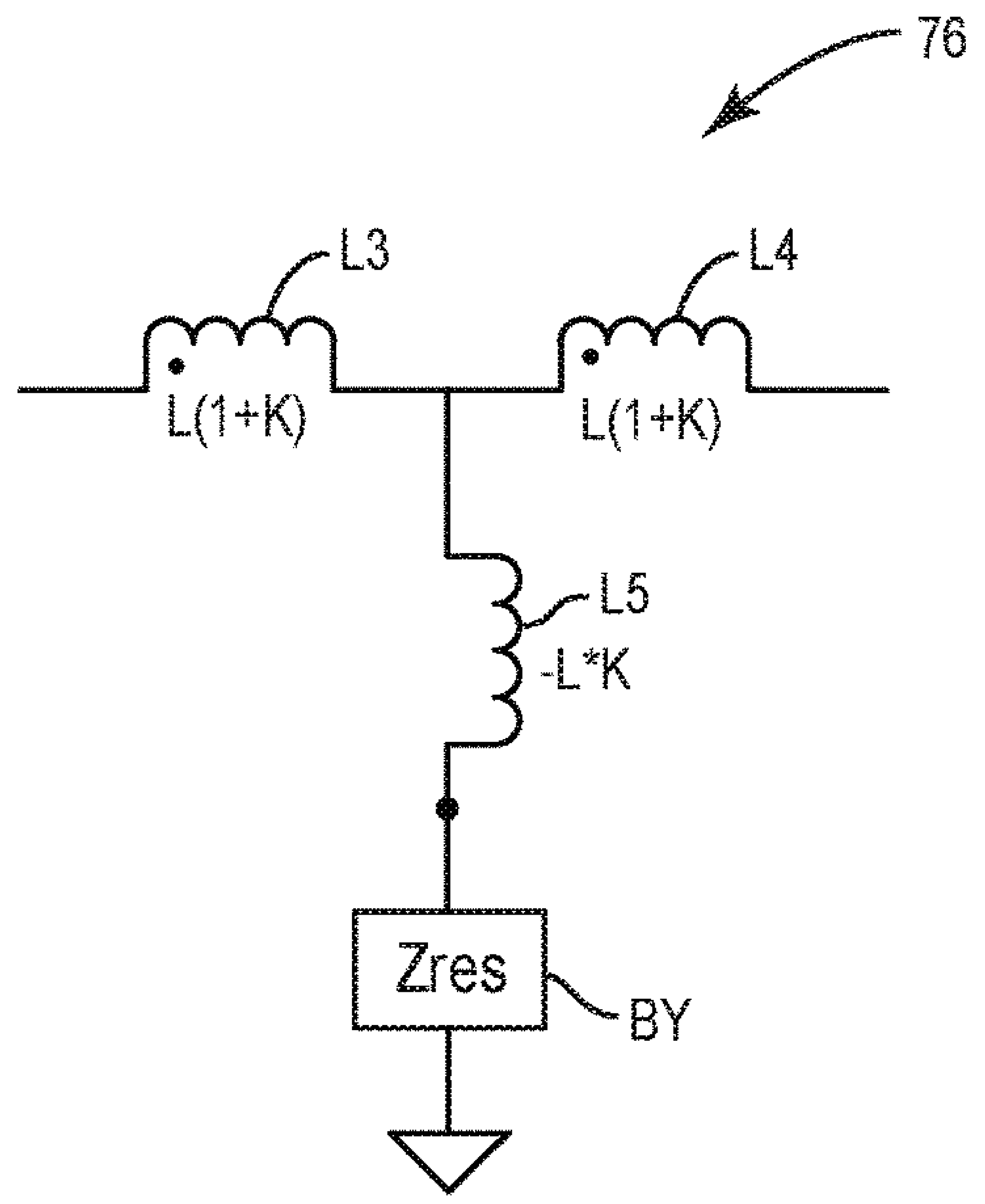
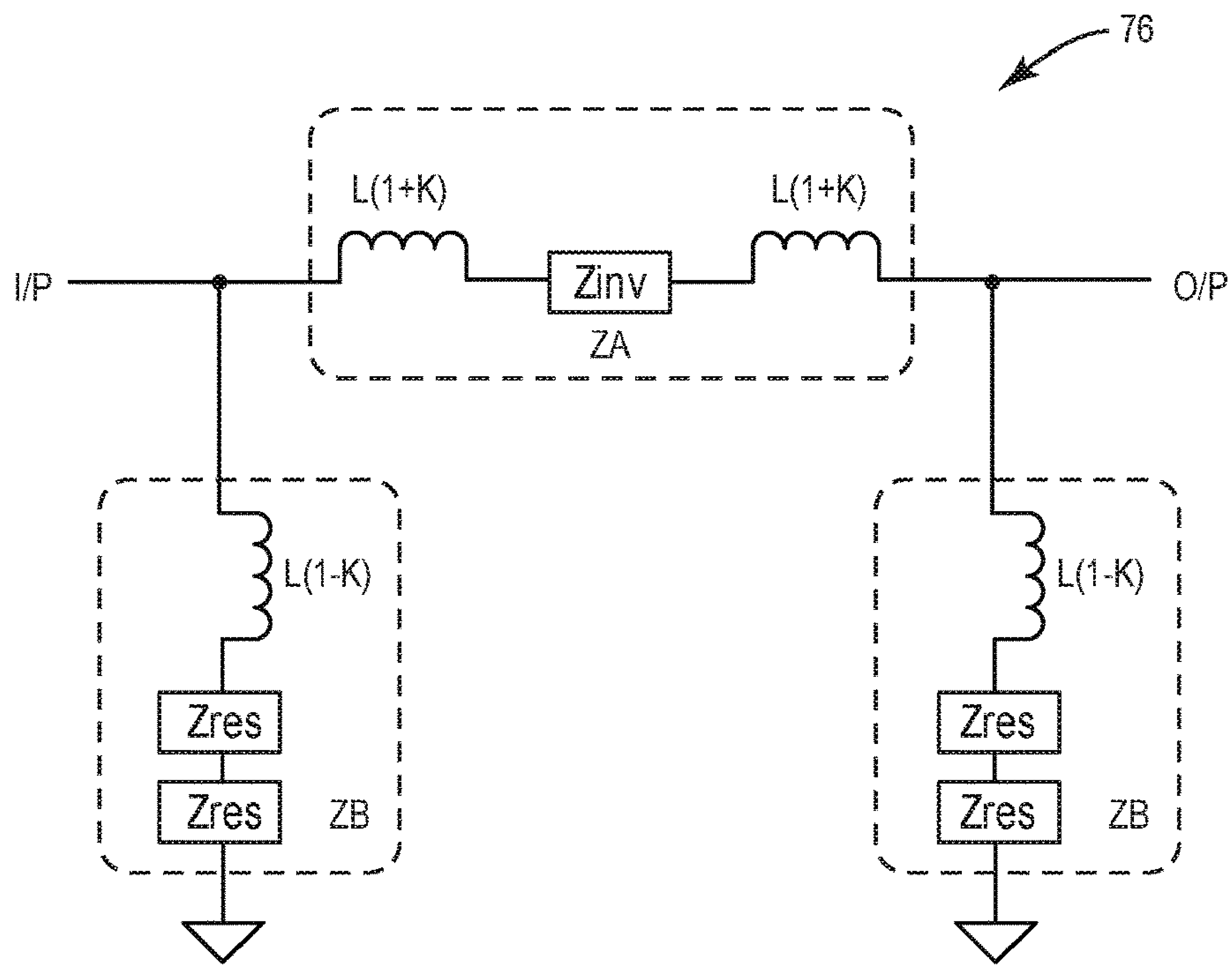


FIG. 19B



T-MODEL OF COMPENSATION CIRCUIT

FIG. 19C



π MODEL OF COMPENSATION CIRCUIT

FIG. 19D

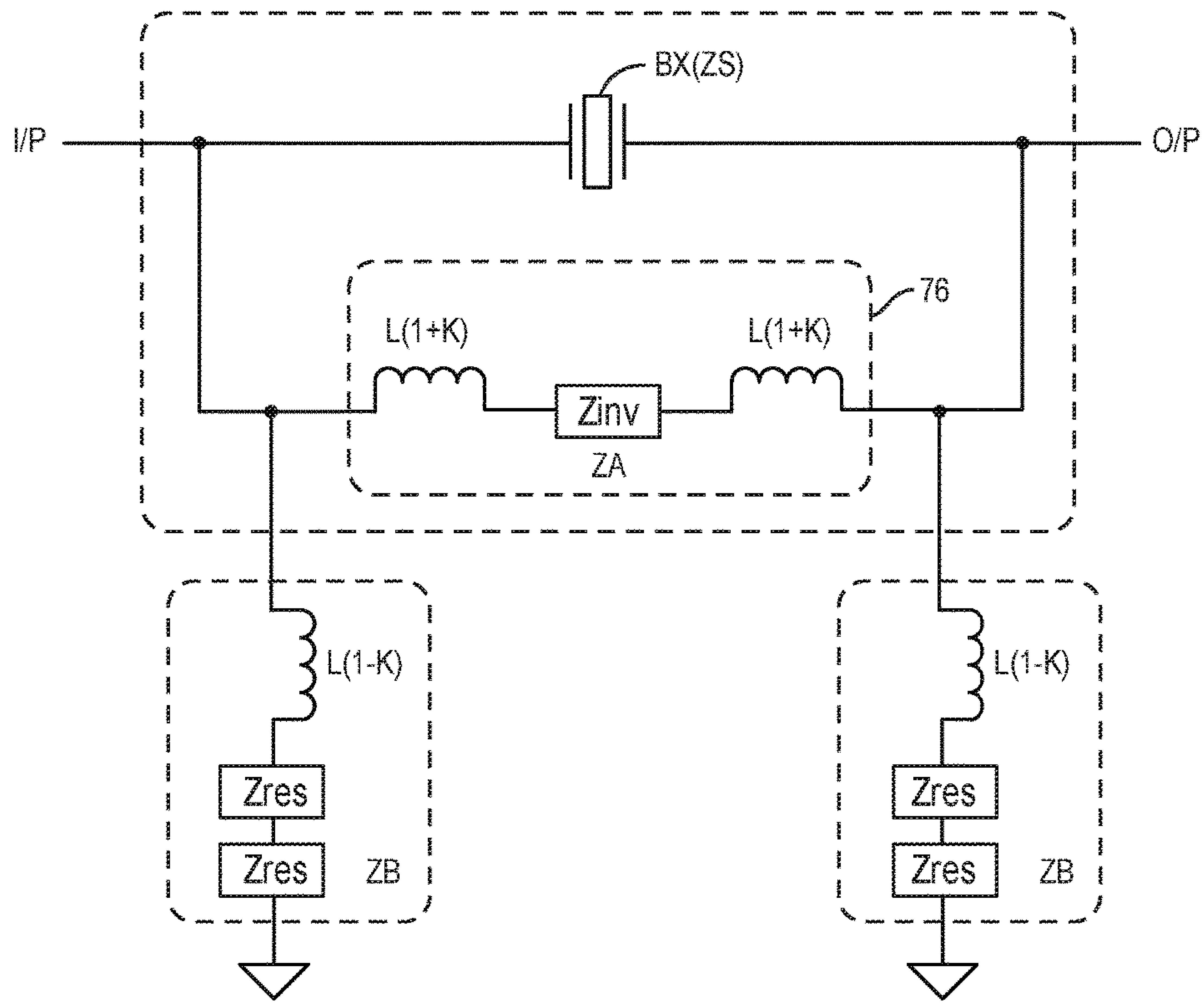


FIG. 20

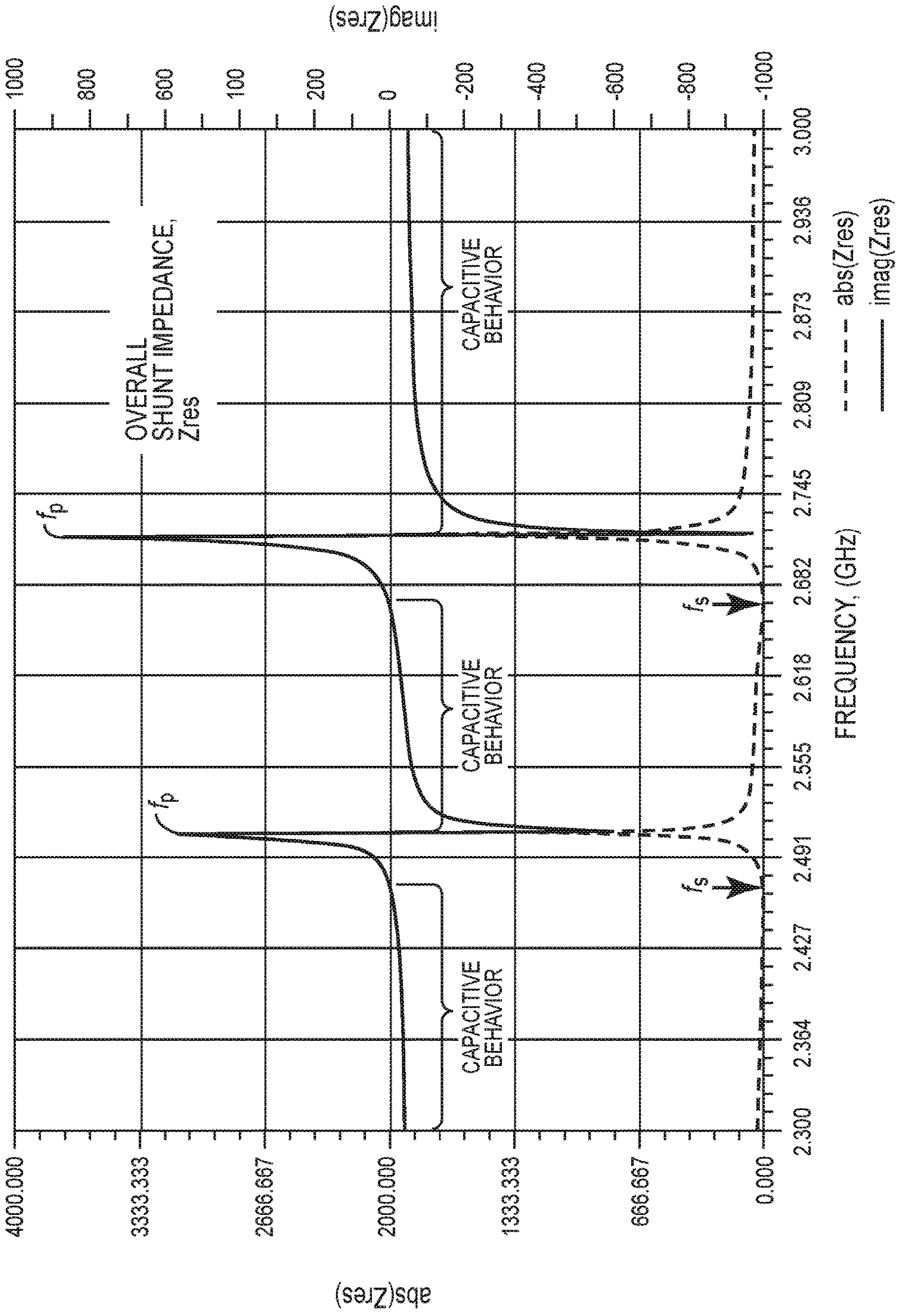

OVERALL SHUNT IMPEDANCE, Zres
imag(Zres)
abs(Zres)
FREQUENCY, (GHz)
CAPACITIVE BEHAVIOR
CAPACITIVE BEHAVIOR
CAPACITIVE BEHAVIOR
f_p
f_p
f_s
f_s
abs(Zres)
imag(Zres)
4000.000
3333.333
2666.667
2000.000
1333.333
666.667
0.000
1000
800
600
100
200
0
-200
-400
-600
-800
-1000
2.300
2.364
2.427
2.491
2.555
2.618
2.682
2.745
2.809
2.873
2.936
3.000

FIG. 21

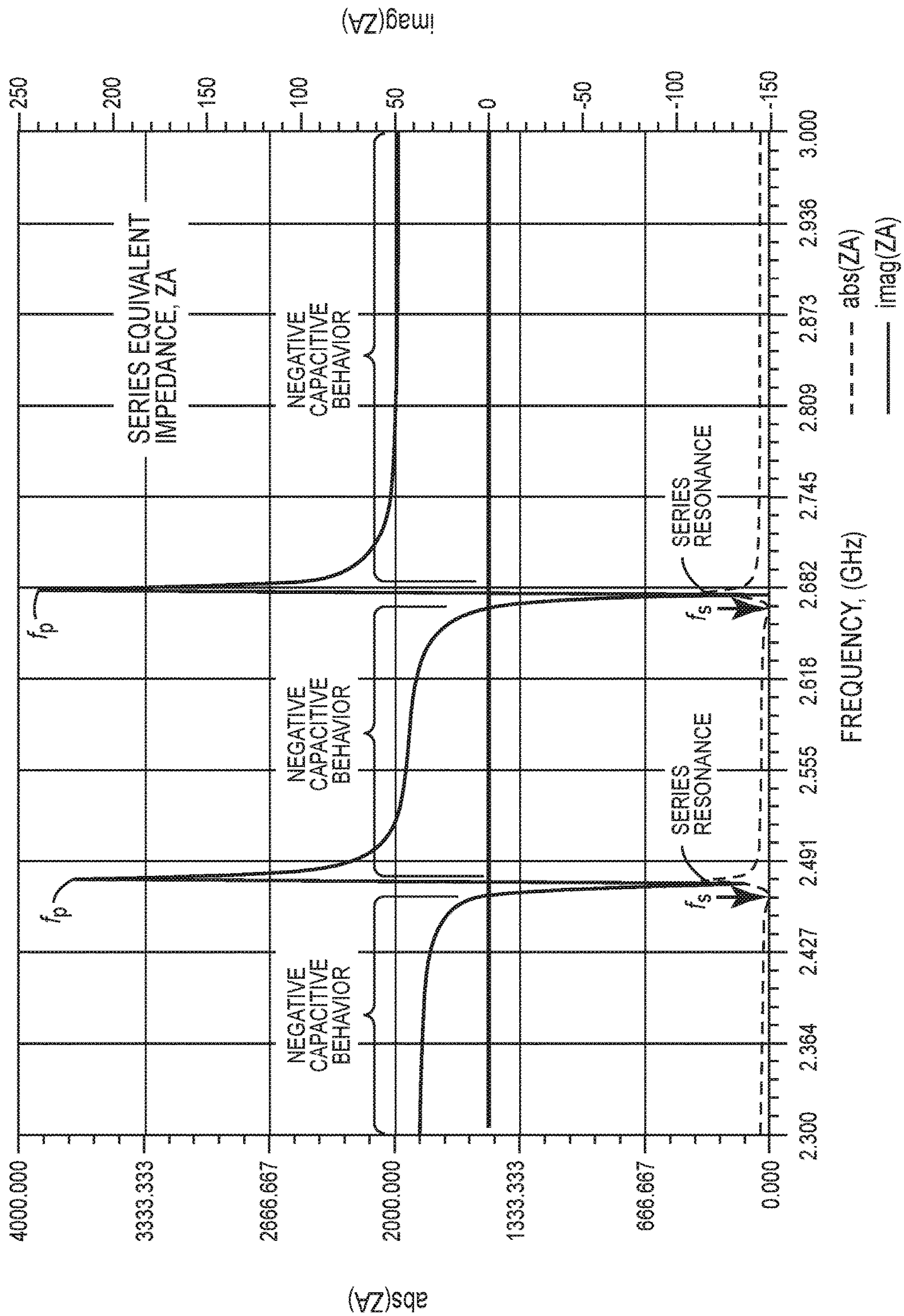

SERIES EQUIVALENT IMPEDANCE, ZA
NEGATIVE CAPACITIVE BEHAVIOR
NEGATIVE CAPACITIVE BEHAVIOR
NEGATIVE CAPACITIVE BEHAVIOR
SERIES RESONANCE
SERIES RESONANCE
f_p
f_p
f_s
f_s
imag(ZA)
250
200
150
100
50
0
-50
-100
-150
abs(ZA)
4000.000
3333.333
2666.667
2000.000
1333.333
666.667
0.000
2.300
2.364
2.427
2.491
2.555
2.618
2.682
2.745
2.809
2.873
2.936
3.000
FREQUENCY, (GHz)
abs(ZA)
imag(ZA)

FIG. 22

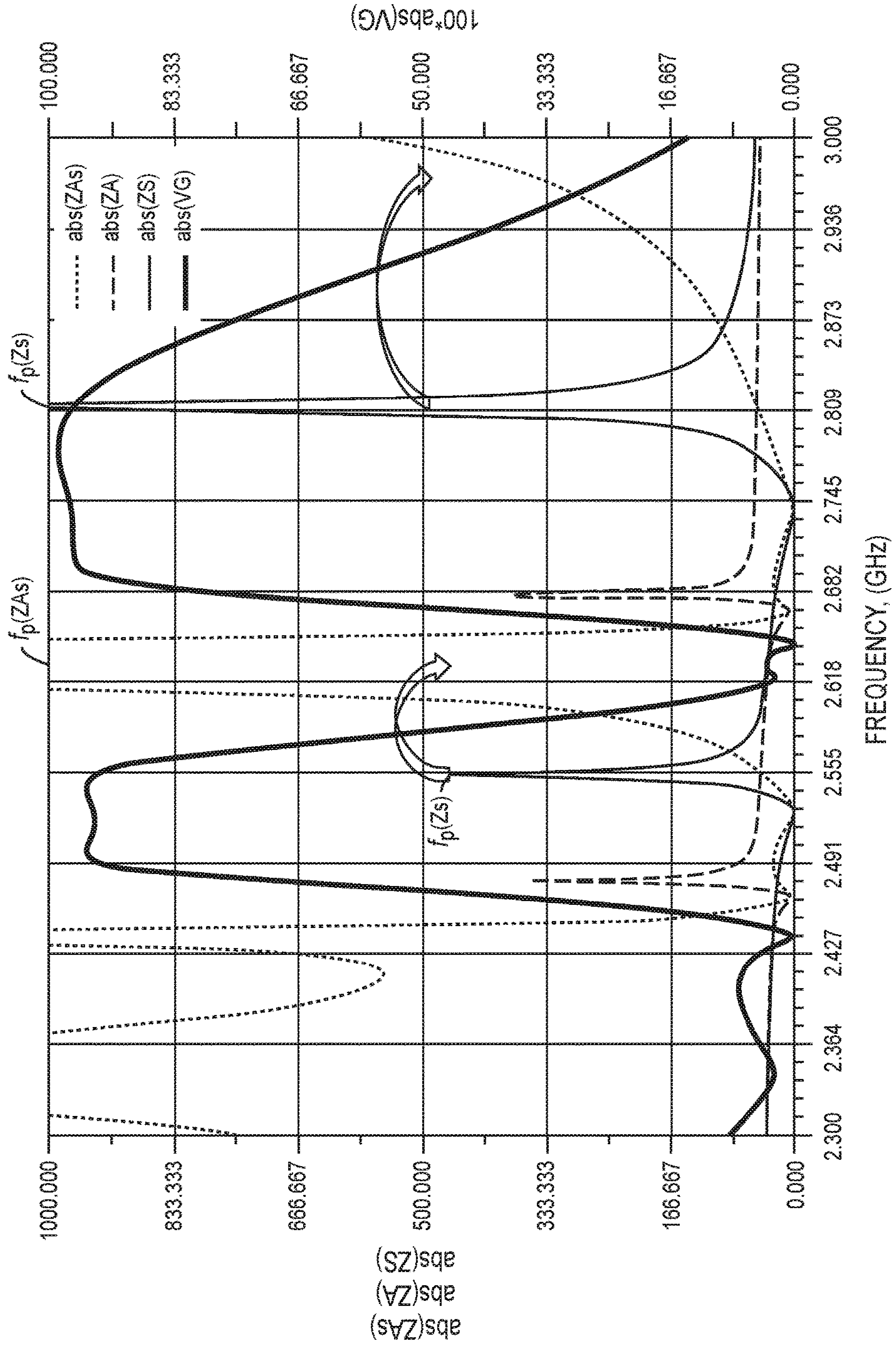
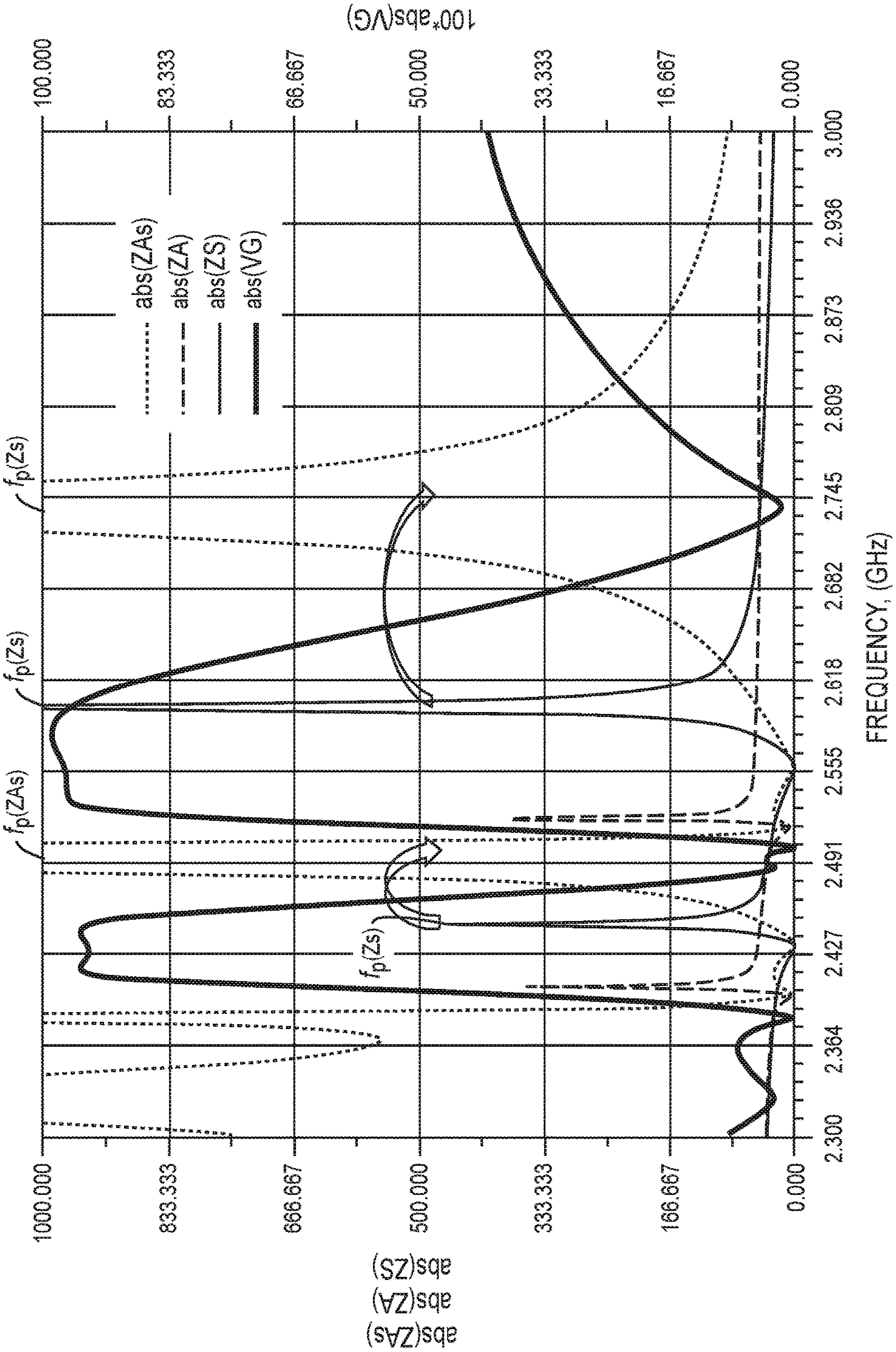

100*abs(VG)
100.000
83.333
66.667
50.000
33.333
16.667
0.000
abs(ZAs)
abs(ZA)
abs(ZS)
1000.000
833.333
666.667
500.000
333.333
166.667
0.000
abs(ZAs)
abs(ZA)
abs(ZS)
abs(VG)
f_p(ZAs)
f_p(Zs)
f_p(Zs)
2.300
2.364
2.427
2.491
2.555
2.618
2.682
2.745
2.809
2.873
2.936
3.000
FREQUENCY, (GHz)

FIG. 23A



100*abs(VG)
100.000
83.333
66.667
50.000
33.333
16.667
0.000
1000.000
833.333
666.667
500.000
333.333
166.667
0.000
abs(ZAs)
abs(ZA)
abs(ZS)
abs(ZAs)
abs(ZA)
abs(ZS)
abs(VG)
f_p(ZAs)
f_p(Zs)
f_p(Zs)
f_p(Zs)
2.300
2.364
2.427
2.491
2.555
2.618
2.682
2.745
2.809
2.873
2.936
3.000
FREQUENCY, (GHz)

FIG. 23B

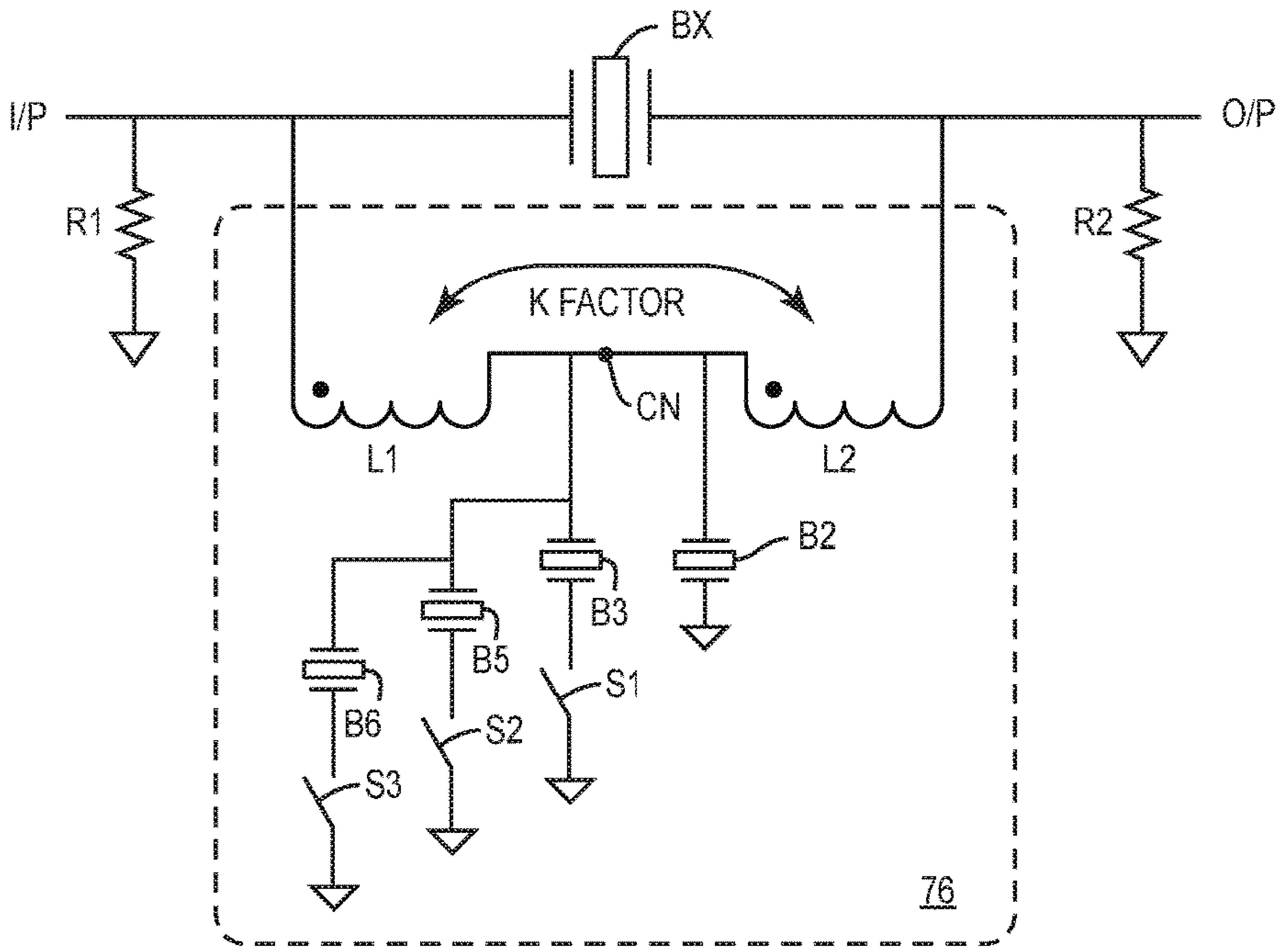

BX
I/P
O/P
R1
R2
K FACTOR
CN
L1
L2
B2
B3
B5
B6
S1
S2
S3
76

FIG. 24

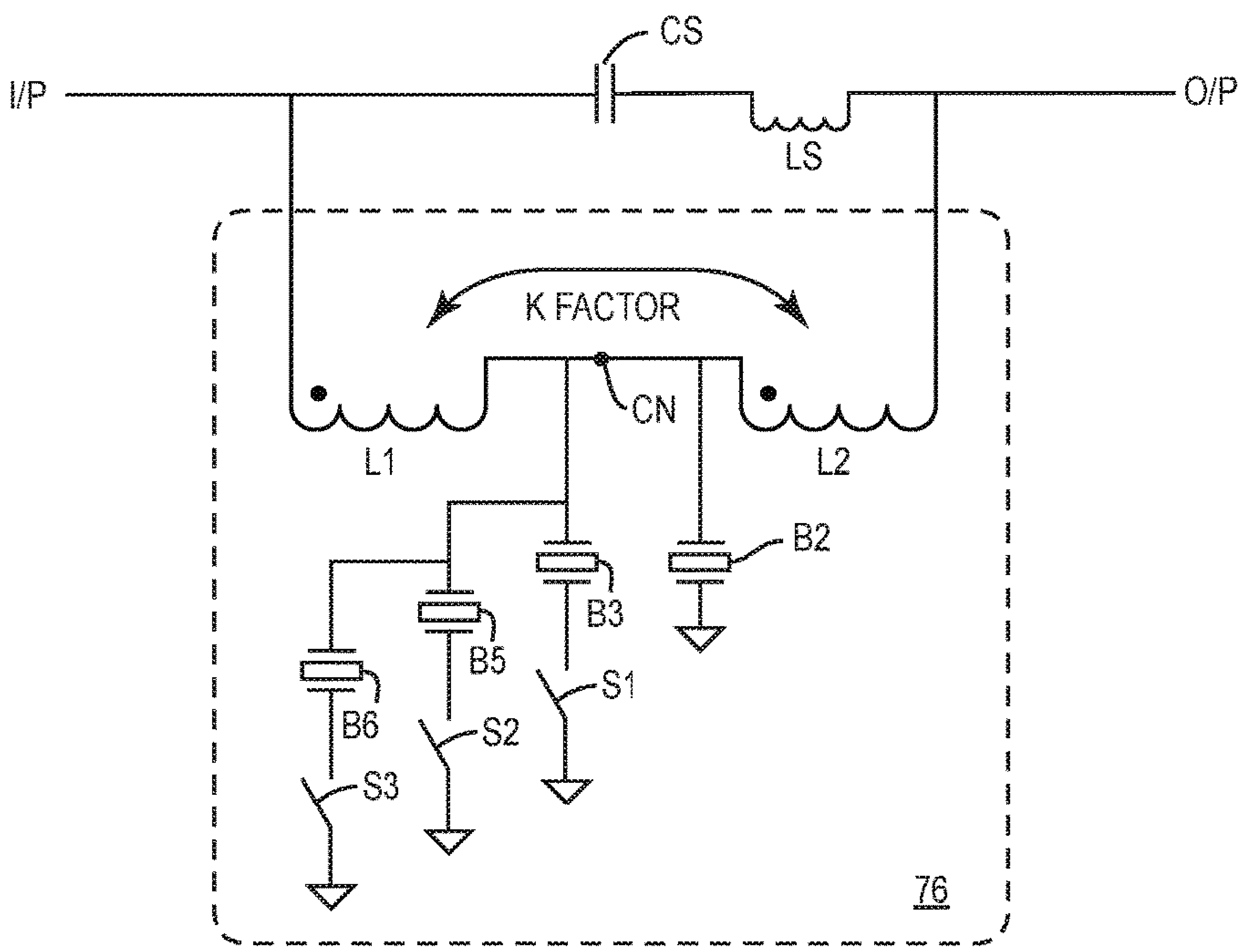

CS
I/P
O/P
LS
K FACTOR
CN
L1
L2
B2
B3
B5
B6
S1
S2
S3
76

FIG. 25

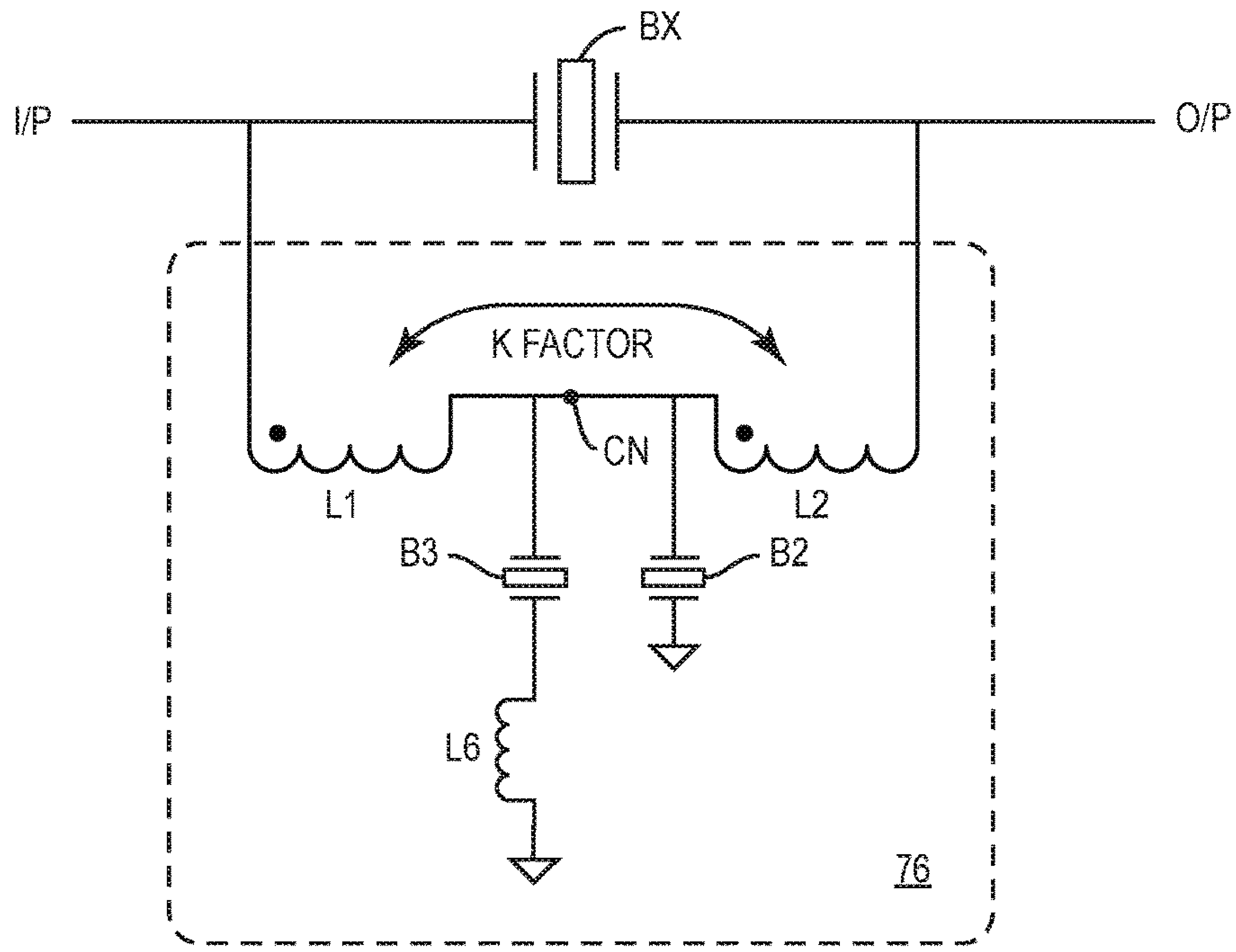

BX
I/P
O/P
K FACTOR
CN
L1
L2
B3
B2
L6
76

FIG. 26

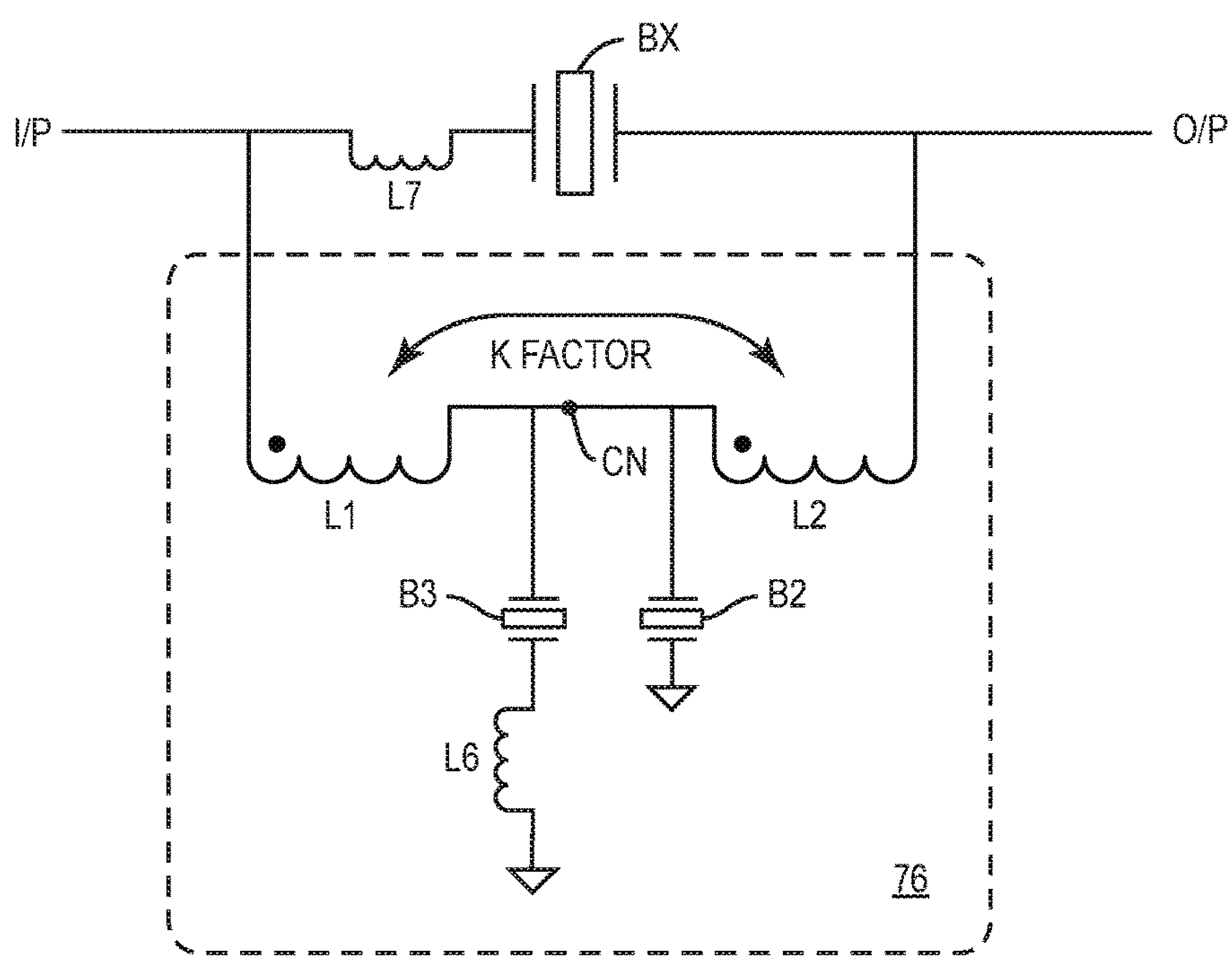

BX
I/P
O/P
L7
K FACTOR
CN
L1
L2
B3
B2
L6
76

FIG. 27

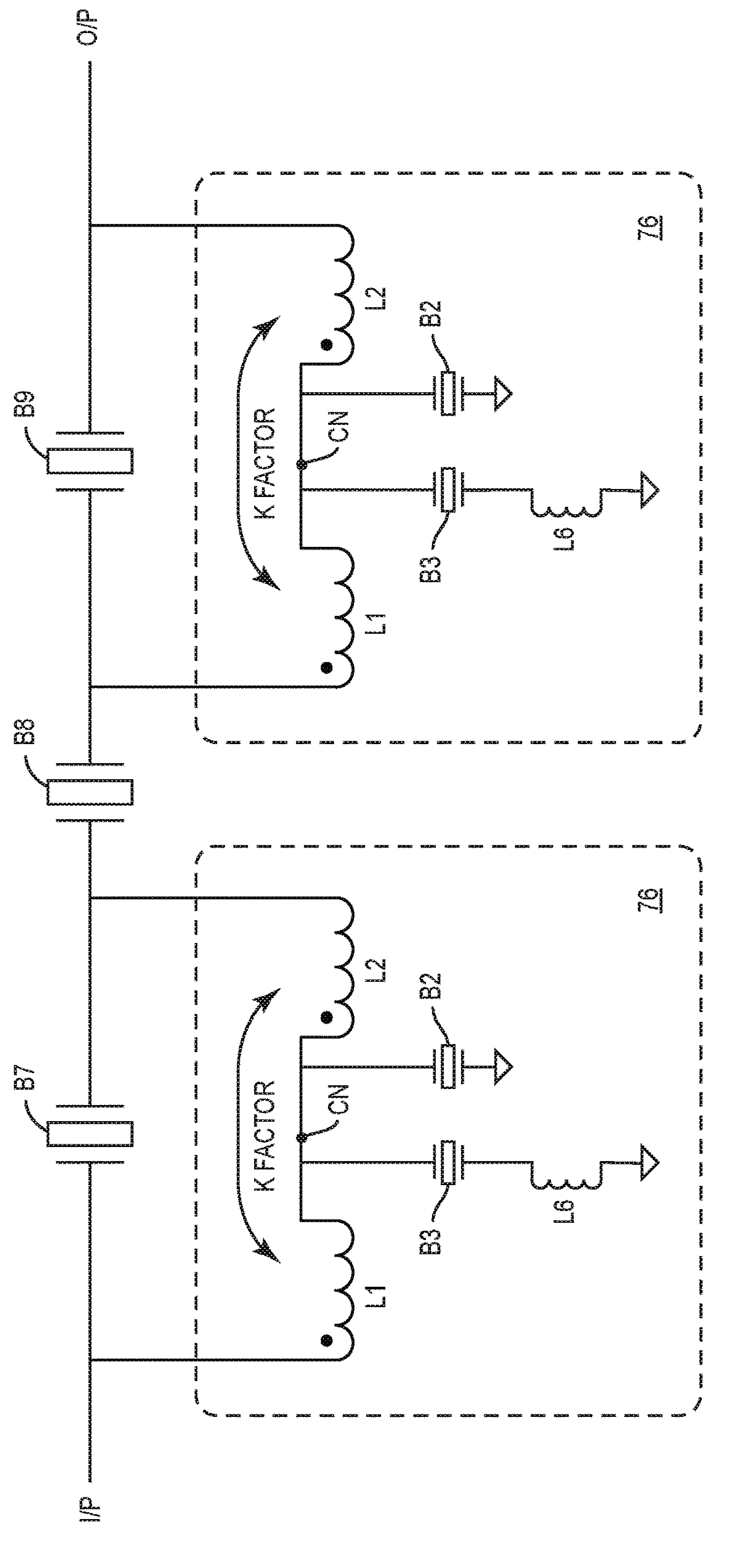

O/P
B9
B8
B7
I/P
K FACTOR
CN
L1
L2
B2
B3
L6
76
K FACTOR
CN
L1
L2
B2
B3
L6
76

FIG. 28

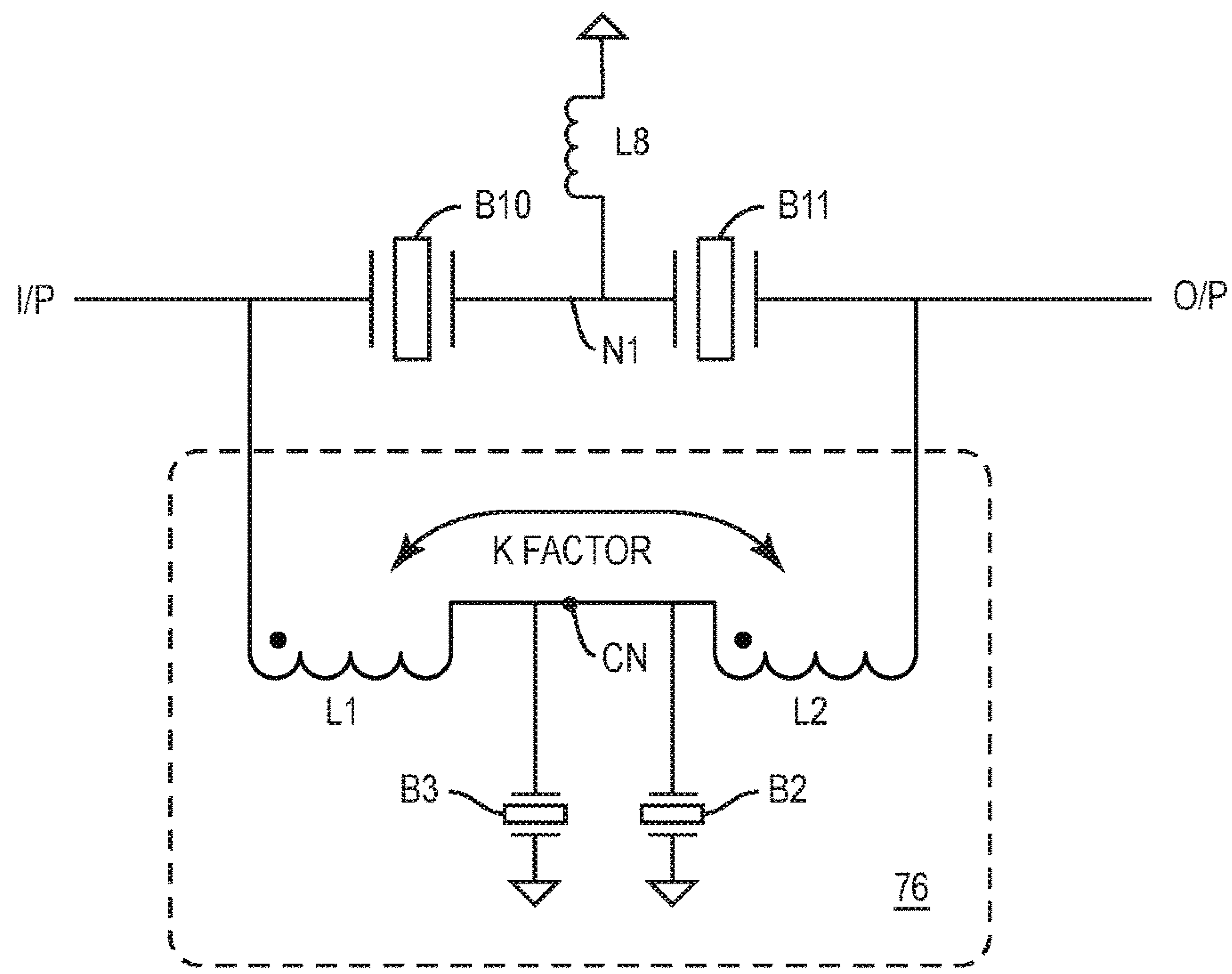

L8
B10
B11
I/P
O/P
N1
K FACTOR
CN
L1
L2
B3
B2
76

FIG. 29

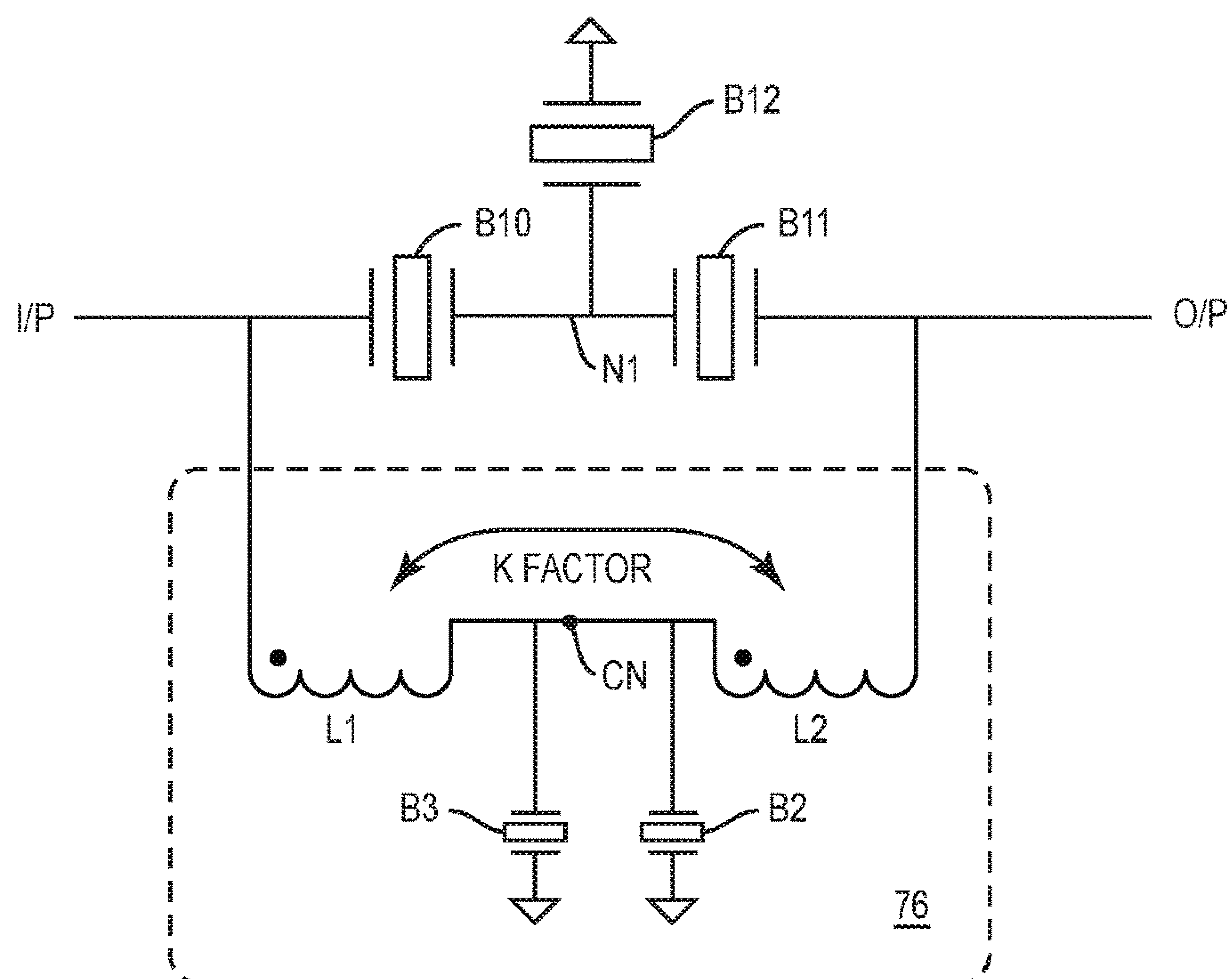

B12
B10
B11
I/P
O/P
N1
K FACTOR
CN
L1
L2
B3
B2
76

FIG. 30

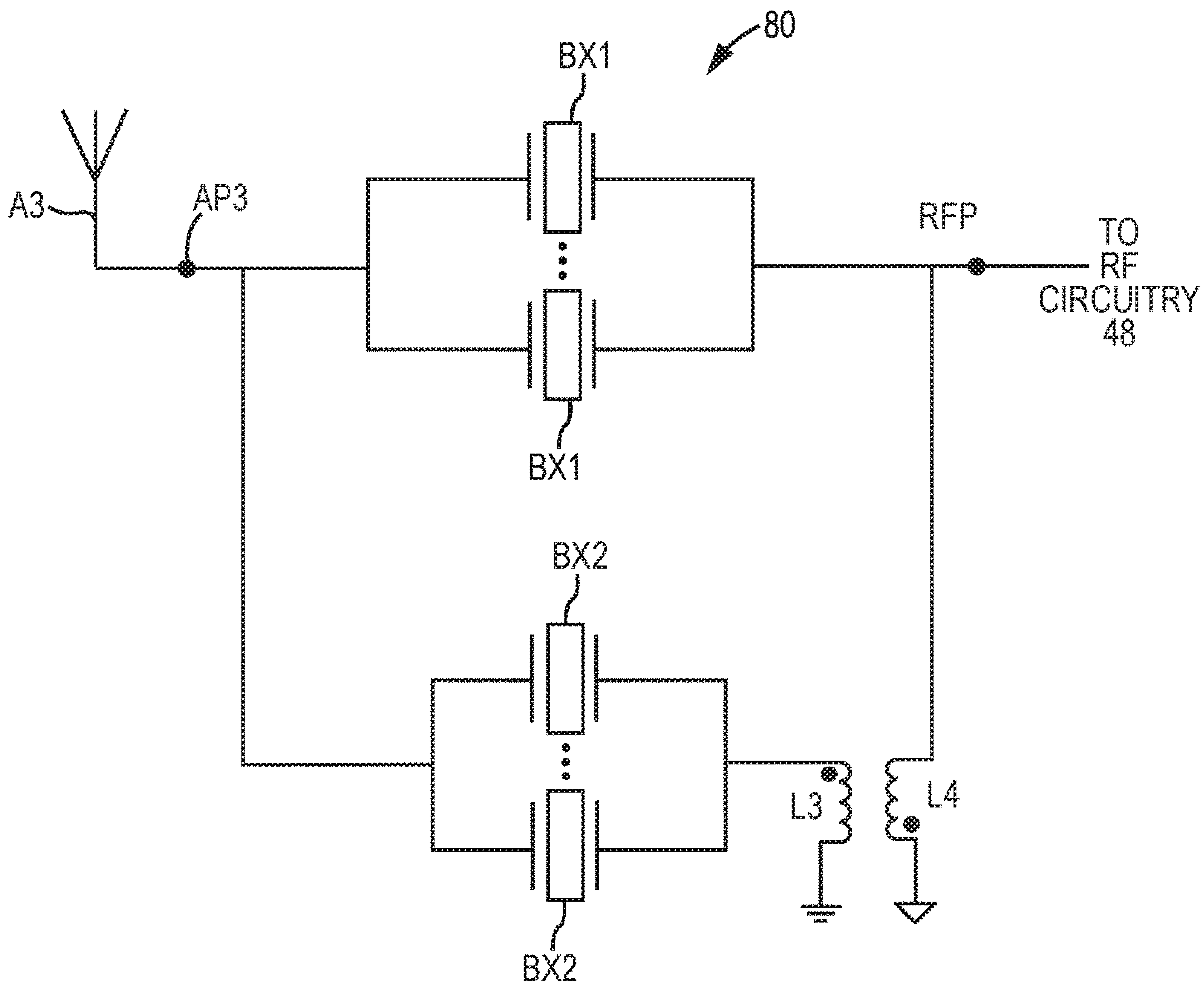


FIG. 31

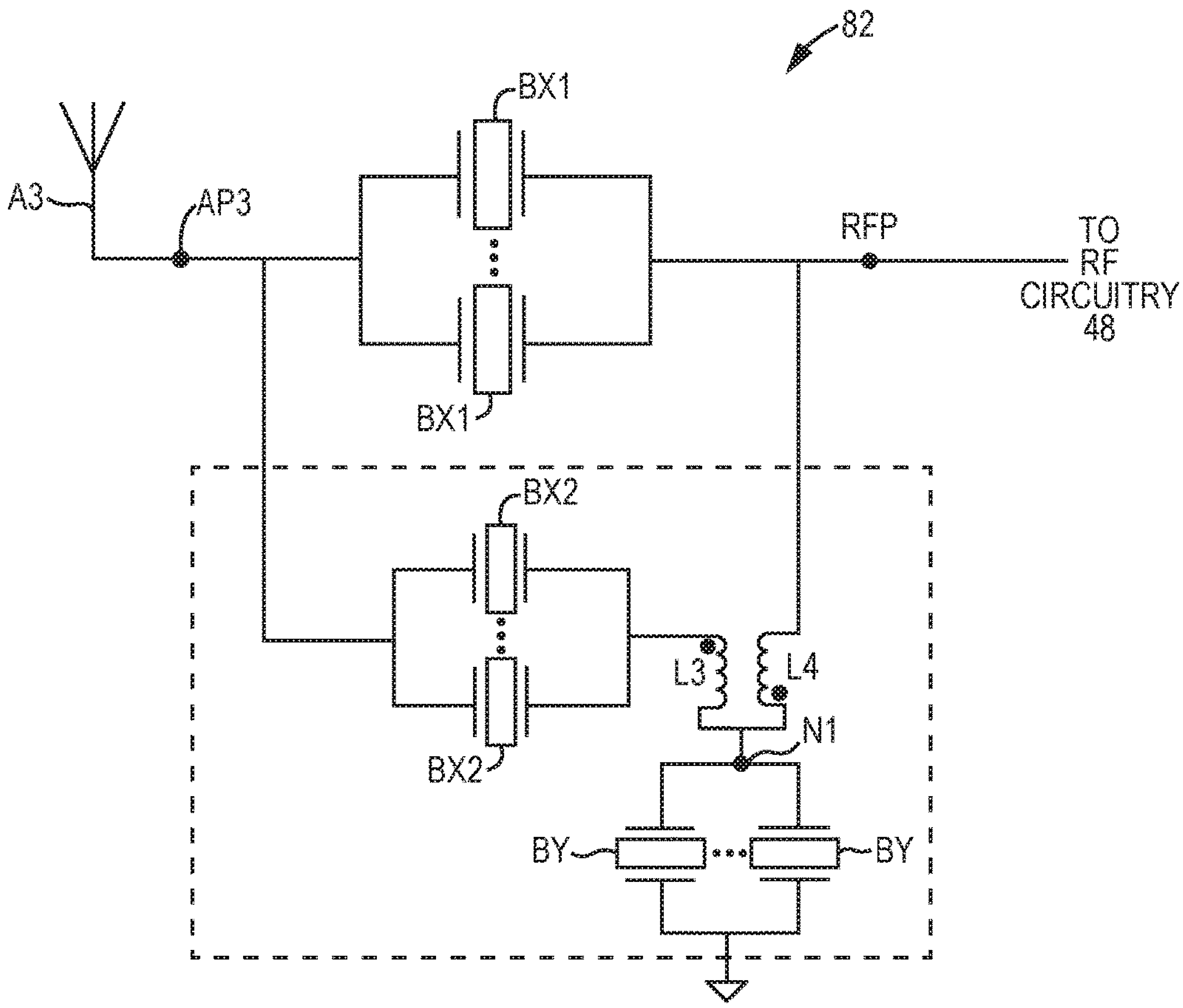


FIG. 32

ACOUSTIC FILTER FOR ANTENNAS

RELATED APPLICATIONS

This application claims the benefit of U.S. provisional patent application Ser. No. 62/324,058, filed Apr. 18, 2016, the disclosure of which is hereby incorporated herein by reference in its entirety.

FIELD OF THE INVENTION

The present disclosure relates to acoustic filters that employ acoustic resonators, and in particular to acoustic filters for antennas.

BACKGROUND

Acoustic resonators, such as Surface Acoustic Wave (SAW) resonators and Bulk Acoustic Wave (BAW) resonators, are used in many high-frequency communication applications. In particular, SAW resonators are often employed in filter networks that operate frequencies up to 1.8 GHz, and BAW resonators are often employed in filter networks that operate at frequencies above 1.5 GHz. Such filters need to have flat passbands, have steep filter skirts and squared shoulders at the upper and lower ends of the passband, and provide excellent rejection outside of the passband. SAW- and BAW-based filters also have relatively low insertion loss, tend to decrease in size as the frequency of operation increases, and are relatively stable over wide temperature ranges. As such, SAW- and BAW-based filters are the filter of choice for many 3rd Generation (3G) and 4th Generation (4G) wireless devices and are destined to dominate filter applications for 5th Generation (5G) wireless devices. Most of these wireless devices support cellular, wireless fidelity (Wi-Fi), Bluetooth, and/or near field communications on the same wireless device and, as such, pose extremely challenging filtering demands. While these demands keep raising the complexity of wireless devices, there is a constant need to improve the performance of acoustic resonators and filters that are based thereon.

To better understand acoustic resonators and various terminology associated therewith, the following provides an overview of a BAW resonator. However, the concepts described herein may employ any type of acoustic resonator and are not limited to SAW- and BAW-based resonators. An exemplary BAW resonator **10** is illustrated in FIG. **1**. The BAW resonator **10** generally includes a substrate **12**, a reflector **14** mounted over the substrate **12**, and a transducer **16** mounted over the reflector **14**. The transducer **16** rests on the reflector **14** and includes a piezoelectric layer **18**, which is sandwiched between a top electrode **20** and a bottom electrode **22**. The top and bottom electrodes **20** and **22** may be formed of Tungsten (W), Molybdenum (Mo), Platinum (Pt), or like material, and the piezoelectric layer **18** may be formed of Aluminum Nitride (AlN), Zinc Oxide (ZnO), or other appropriate piezoelectric material. Although shown in FIG. **1** as each including a single layer, the piezoelectric layer **18**, the top electrode **20**, and/or the bottom electrode **22** may include multiple layers of the same material, multiple layers in which at least two layers are different materials, or multiple layers in which each layer is a different material.

The BAW resonator **10** is divided into an active region **24** and an outside region **26**. The active region **24** generally corresponds to the section of the BAW resonator **10** where the top and bottom electrodes **20** and **22** overlap and also includes the layers below the overlapping top and bottom

electrodes **20** and **22**. The outside region **26** corresponds to the section of the BAW resonator **10** that surrounds the active region **24**.

For the BAW resonator **10**, applying electrical signals across the top electrode **20** and the bottom electrode **22** excites acoustic waves in the piezoelectric layer **18**. These acoustic waves primarily propagate vertically. A primary goal in BAW resonator design is to confine these vertically propagating acoustic waves in the transducer **16**. Acoustic waves traveling upward are reflected back into the transducer **16** by the air-metal boundary at the top surface of the top electrode **20**. Acoustic waves traveling downward are reflected back into the transducer **16** by the reflector **14** or by an air cavity, which is provided just below the transducer in a Film BAW Resonator (FBAR).

The reflector **14** is typically formed by a stack of reflector layers (RL) **28**, which alternate in material composition to produce a significant reflection coefficient at the junction of adjacent reflector layers **28**. Typically, the reflector layers **28** alternate between materials having high and low acoustic impedances, such as tungsten (W) and silicon dioxide (SiO_2). While only five reflector layers **28** are illustrated in FIG. **1**, the number of reflector layers **28** and the structure of the reflector **14** varies from one design to another.

The magnitude (Z) and phase (ϕ) of the electrical impedance as a function of the frequency for a relatively ideal BAW resonator **10** is provided in FIG. **2**. The magnitude (Z) of the electrical impedance is illustrated by the solid line, whereas the phase (ϕ) of the electrical impedance is illustrated by the dashed line. A unique feature of the BAW resonator **10** is that it has both a resonance frequency and an anti-resonance frequency. The resonance frequency is typically referred to as the series resonance frequency (f_s), the anti-resonance frequency is typically referred to as the parallel resonance frequency (f_p). The series resonance frequency (f_s) occurs when the magnitude of the impedance, or reactance, of the BAW resonator **10** approaches zero. The parallel resonance frequency (f_p) occurs when the magnitude of the impedance, or reactance, of the BAW resonator **10** peaks at a significantly high level. In general, the series resonance frequency (f_s) is a function of the thickness of the piezoelectric layer **18** and the mass of the bottom and top electrodes **20** and **22**.

For the phase, the BAW resonator **10** acts like an inductance that provides a 90° phase shift between the series resonance frequency (f_s) and the parallel resonance frequency (f_p). In contrast, the BAW resonator **10** acts like a capacitance that provides a −90° phase shift below the series resonance frequency (f_s) and above the parallel resonance frequency (f_p). The BAW resonator **10** presents a very low, near zero, resistance at the series resonance frequency (f_s) and a very high resistance at the parallel resonance frequency (f_p). The electrical nature of the BAW resonator **10** lends itself to the realization of a very high Q (quality factor) inductance over a relatively short range of frequencies, which has proved to be very beneficial in high-frequency filter networks, especially those operating at frequencies around 1.8 GHz and above.

Unfortunately, the phase (ϕ) curve of FIG. **2** is representative of an ideal phase curve. In reality, approaching this ideal is challenging. A typical phase curve for the BAW resonator **10** of FIG. **1** is illustrated in FIG. **3**A. Instead of being a smooth curve, the phase curve of FIG. **3**A includes ripple below the series resonance frequency (f_s), between the series resonance frequency (f_s) and the parallel resonance frequency (f_p), and above the parallel resonance frequency (f_p). The ripple is the result of spurious modes, which are caused by spurious resonances that occur in corresponding frequencies. While the vast majority of the acoustic waves in the BAW resonator **10** propagate vertically, various boundary conditions about the transducer **16** result in the propagation of lateral (horizontal) acoustic waves, which are referred to as lateral standing waves. The presence of these lateral standing waves reduces the potential Q associated with the BAW resonator **10**.

As illustrated in FIG. **4**, a border (BO) ring **30** is formed on or within the top electrode **20** to suppress certain of the spurious modes. The spurious modes that are suppressed by the BO ring **30** are those above the series resonance frequency (f_s), as highlighted by circles A and B in the phase curve of FIG. **3**B. Circle A shows a suppression of the ripple, and thus of the spurious mode, in the passband of the phase curve, which resides between the series resonance frequency (f_s) and the parallel resonance frequency (f_p). Circle B shows suppression of the ripple, and thus of the spurious modes, above the parallel resonance frequency (f_p). Notably, the spurious mode in the upper shoulder of the passband, which is just below the parallel resonance frequency f_p, and the spurious modes above the passband are suppressed, as evidenced by the smooth or substantially ripple free phase curve between the series resonance frequency (f_s) and the parallel resonance frequency (f_p) and above the parallel resonance frequency (f_p).

The BO ring **30** corresponds to a mass loading of the portion of the top electrode **20** that extends about the periphery of the active region **24**. The BO ring **30** may correspond to a thickened portion of the top electrode **20** or the application of additional layers of an appropriate material over the top electrode **20**. The portion of the BAW resonator **10** that includes and resides below the BO ring **30** is referred to as a BO region **32**. Accordingly, the BO region **32** corresponds to an outer, perimeter portion of the active region **24** and resides inside of the active region **24**.

While the BO ring **30** is effective at suppressing spurious modes above the series resonance frequency (f_s), the BO ring **30** has little or no impact on those spurious modes below the series resonance frequency (f_s), as shown by the ripples in the phase curve below the series resonance frequency (f_s) in FIG. **3**B. A technique referred to as apodization is often used to suppress the spurious modes that fall below the series resonance frequency (f_s).

Apodization tries to avoid, or at least significantly reduce, any lateral symmetry in the BAW resonator **10**, or at least in the transducer **16** thereof. The lateral symmetry corresponds to the footprint of the transducer **16**, and avoiding the lateral symmetry corresponds to avoiding symmetry associated with the sides of the footprint. For example, one may choose a footprint that corresponds to a pentagon instead of a square or rectangle. Avoiding symmetry helps reduce the presence of lateral standing waves in the transducer **16**. Circle C of FIG. **3**C illustrates the effect of apodization in which the spurious modes below the series resonance frequency (f_s) are suppressed, as evidence by the smooth or substantially ripple free phase curve below the series resonance frequency (f_s). Assuming no BO ring **30** is provided, one can readily see in FIG. **3**C that apodization fails to suppress those spurious modes above the series resonance frequency (f_s). As such, the typical BAW resonator **10** employs both apodization and the BO ring **30**.

As noted previously, BAW resonators **10** are often used in filter networks that operate at high frequencies and require high Q values. A basic ladder network **40** is illustrated in FIG. **5**A. The ladder network **40** includes two series resonators B_{SER} and two shunt resonators B_{SH}, which are arranged in a traditional ladder configuration. Typically, the series resonators B_{SER} have the same or similar first frequency response, and the shunt resonators B_{SH} have the same or similar second frequency response, which is different from the first frequency response, as shown in FIG. **5**B. In many applications, the shunt resonators B_{SH} are detuned versions of the series resonators B_{SER}. As a result, the frequency responses for the series resonators B_{SER} and the shunt resonators B_{SH} are generally very similar, yet shifted relative to one another such that the parallel resonance frequency ($f_{p,SH}$) of the shunt resonators approximates the series resonance frequency ($f_{s,SER}$) of the series resonators B_{SER}. Note that the series resonance frequency ($f_{s,SH}$) of the shunt resonators B_{SH} is less than the series resonance frequency ($f_{s,SER}$) of the series resonators B_{SER}. The parallel resonance frequency ($f_{p,SH}$) of the shunt resonators B_{SH} is less than the parallel resonance frequency ($f_{p,SER}$) of the series resonators B_{SER}.

FIG. **5**C is associated with FIG. **5**B and illustrates the response of the ladder network **40**. The series resonance frequency ($f_{s,SH}$) of the shunt resonators B_{SH} corresponds to the low side of the passband's skirt (phase **2**), and the parallel resonance frequency ($f_{p,SER}$) of the series resonators B_{SER} corresponds to the high side of the passband's skirt (phase **4**). The substantially aligned series resonance frequency ($f_{s,SER}$) of the series resonators B_{SER} and the parallel resonance frequency ($f_{p,SH}$) of the shunt resonators B_{SH} fall within the passband. FIGS. **6**A through **6**E provide circuit equivalents for the five phases of the response of the ladder network **40**. During the first phase (phase **1**, FIGS. **5**C, **6**A), the ladder network **40** functions to attenuate the input signal. As the series resonance frequency ($f_{s,SH}$) of the shunt resonators B_{SH} is approached, the impedance of the shunt resonators B_{SH} drops precipitously such that the shunt resonators B_{SH} essentially provide a short to ground at the series resonance frequency ($f_{s,SH}$) of the shunt resonators (phase **2**, FIGS. **5**C, **6**B). At the series resonance frequency ($f_{s,SH}$) of the shunt resonators B_{SH} (phase **2**), the input signal is essentially blocked from the output of the ladder network **40**.

Between the series resonance frequency ($f_{s,SH}$) of the shunt resonators B_{SH} and the parallel resonance frequency ($f_{p,SER}$) of the series resonators B_{SER}, which corresponds to the passband, the input signal is passed to the output with relatively little or no attenuation (phase **3**, FIGS. **5**C, **6**C). Within the passband, the series resonators B_{SER} present relatively low impedance, whereas the shunt resonators B_{SH} present a relatively high impedance, wherein the combination of the two leads to a flat passband with steep low- and high-side skirts. As the parallel resonance frequency ($f_{p,SER}$) of the series resonators B_{SER} is approached, the impedance of the series resonators B_{SER} becomes very high, such that the series resonators B_{SER} essentially present themselves as open at the parallel resonance frequency ($f_{p,SER}$) of the series resonators (phase **4**, FIGS. **5**C, **6**D). At the parallel resonance frequency ($f_{p,SER}$) of the series resonators B_{SER} (phase **4**), the input signal is again essentially blocked from the output of the ladder network **40**. During the final phase (phase **5**, FIGS. **5**C, **6**E), the ladder network **40** functions to attenuate the input signal, in a similar fashion to that provided in phase **1**. As the parallel resonance frequency ($f_{p,SER}$) of the series resonators B_{SER} is passed, the impedance of the series resonators B_{SER} decreases and the impedance of the shunt resonators B_{SH} normalizes. Thus, the ladder network **40** functions to provide a high Q passband between the series resonance frequency ($f_{s,SH}$) of the shunt resonators B_{SH} and the parallel resonance frequency ($f_{p,SER}$) of the series resonators B_{SER}. The ladder network **40** provides extremely high attenuation at both the series resonance frequency ($f_{s,SH}$) of the shunt resonators B_{SH} and the parallel resonance frequency ($f_{p,SER}$) of the series resonators. The ladder network **40** provides good attenuation below the series resonance frequency ($f_{s,SH}$) of the shunt resonators B_{SH} and above the parallel resonance frequency ($f_{p,SER}$) of the series resonators B_{SER}. As noted previously, there is a constant need to improve the performance of acoustic resonators and filters that are based thereon.

SUMMARY

The present disclosure relates to filter circuitry for use in communication systems that employ multiple antennas. In general, a communication system may have a transmit path and a receive path. The transmit path extends to a first antenna port and is configured to present signals for transmission in a first communication band and a second communication band to the first antenna port for transmission via a first antenna that is coupled to the first antenna port. The receive path extends to a second antenna port and comprises a first multiple passband/multiple stopband (MPMS) filter that provides a plurality of passbands and a plurality of stopbands interleaved with one another, wherein a first stopband and a second stopband of the plurality of stopbands correspond respectively to the first communication band and the second communication band and are separated by a first passband of the plurality of passbands. In one embodiment, the first passband of the plurality of passbands corresponds to a third communication band, and the receive path is configured to receive signals via a second antenna coupled to the second antenna port. The filter circuitry may employ acoustic resonators, such as SAW resonators and BAW resonators.

Those skilled in the art will appreciate the scope of the present disclosure and realize additional aspects thereof after reading the following detailed description of the preferred embodiments in association with the accompanying drawing figures.

BRIEF DESCRIPTION OF THE DRAWING FIGURES

The accompanying drawing figures incorporated in and forming a part of this specification illustrate several aspects of the disclosure and, together with the description, serve to explain the principles of the disclosure.

FIG. **1** illustrates a conventional Bulk Acoustic Wave (BAW) resonator.

FIG. **2** is a graph of the magnitude and phase of impedance over frequency responses as a function of frequency for an ideal BAW resonator.

FIGS. **3**A-**3**C are graphs of phase responses for various BAW resonator configurations.

FIG. **4** illustrates a conventional BAW resonator with a border ring.

FIG. **5**A is a schematic of a conventional ladder network.

FIGS. **5**B and **5**C are graphs of a frequency response for BAW resonators in the conventional ladder network of FIG. **5**A and a frequency response for the conventional ladder network of FIG. **5**A.

FIGS. **6**A-**6**E are circuit equivalents for the ladder network of FIG. **5**A at the frequency points **1**, **2**, **3**, **4**, and **5**, which are identified in FIG. **5**C.

FIG. **7** is a block diagram of a mobile terminal according to one embodiment.

FIG. **8** is a schematic of an RF front-end according to a first embodiment.

FIG. **9** illustrates a rigid PCB and a flexible PCB coupled together by multiple coaxial cables according to a first embodiment.

FIG. **10** is a schematic of an RF front-end according to a second embodiment.

FIG. **11** illustrates an acoustic resonator in parallel with a compensation circuit, which includes a single shunt acoustic resonator.

FIG. **12** is a graph that illustrates exemplary frequency responses for the acoustic resonator, compensation circuit, and overall circuit of FIG. **11**.

FIG. **13** illustrates an acoustic resonator in parallel with a compensation circuit, which includes at least two shunt acoustic resonators, according to a first embodiment.

FIG. **14** is a graph that illustrates exemplary frequency responses for the acoustic resonator, compensation circuit, and overall circuit of FIG. **13**.

FIG. **15** is a graph that compares actual frequency responses of the overall circuits of FIGS. **11** and **13**.

FIG. **16** illustrates a plurality of parallel acoustic resonators in parallel with a compensation circuit, which includes at least two shunt acoustic resonators, according to a second embodiment.

FIG. **17** is a graph that illustrates first exemplary frequency responses for the acoustic resonator, compensation circuit, and overall circuit of FIG. **16**.

FIG. **18** is a graph that illustrates second exemplary frequency responses for the acoustic resonator, compensation circuit, and overall circuit of FIG. **16**.

FIGS. **19**A through **19**D illustrate transformation of the T-circuit impedance architecture of the compensation circuit of FIG. **13** to a π (pi) impedance model.

FIG. **20** illustrates the overall circuit of FIG. **13** using the π (pi) impedance model of FIG. **19**D.

FIG. **21** is a graph illustrating the overall shunt impedance, Zres, according to one embodiment.

FIG. **22** is a graph illustrating the series equivalent impedance, ZA, according to one embodiment.

FIGS. **23**A and **23**B are graphs over different frequency ranges illustrating the absolute or magnitude of series impedance ZS, the series equivalent impedance ZA, and overall series impedance ZAs, according to one embodiment.

FIG. **24** illustrates an acoustic resonator in parallel with a compensation circuit, which includes at least two shunt acoustic resonators, according to a third embodiment.

FIG. **25** illustrates a series resonant inductor-capacitor (L-C) circuit in parallel with a compensation circuit, which includes at least two shunt acoustic resonators, according to a fourth embodiment.

FIG. **26** illustrates an acoustic resonator in parallel with a compensation circuit, which includes at least two shunt acoustic resonators, according to a fifth embodiment.

FIG. **27** illustrates an acoustic resonator in parallel with a compensation circuit, which includes at least two shunt acoustic resonators, according to a sixth embodiment.

FIG. **28** illustrates an acoustic resonator in parallel with a compensation circuit, which includes at least two shunt acoustic resonators, according to a seventh embodiment.

FIG. **29** illustrates two series acoustic resonators in parallel with a compensation circuit, which includes at least two shunt acoustic resonators, according to an eighth embodiment.

FIG. **30** illustrates two series acoustic resonators in parallel with a compensation circuit, which includes at least two shunt acoustic resonators, according to a ninth embodiment.

FIG. **31** illustrates a first alternative filter structure that provides multiple passbands and multiple stopbands that are interleaved with one another.

FIG. **32** illustrates a first alternative filter structure that provides multiple passbands and multiple stopbands that are interleaved with one another.

DETAILED DESCRIPTION

The embodiments set forth below represent the necessary information to enable those skilled in the art to practice the embodiments and illustrate the best mode of practicing the embodiments. Upon reading the following description in light of the accompanying drawing figures, those skilled in the art will understand the concepts of the disclosure and will recognize applications of these concepts not particularly addressed herein. It should be understood that these concepts and applications fall within the scope of the disclosure and the accompanying claims.

It will be understood that, although the terms first, second, etc. may be used herein to describe various elements, these elements should not be limited by these terms. These terms are only used to distinguish one element from another. For example, a first element could be termed a second element, and similarly, a second element could be termed a first element, without departing from the scope of the present disclosure. As used herein, the term "and/or" includes any and all combinations of one or more of the associated listed items.

It will be understood that when an element such as a layer, region, or substrate is referred to as being "on" or extending "onto" another element, it can be directly on or extend directly onto the other element or intervening elements may also be present. In contrast, when an element is referred to as being "directly on" or extending "directly onto" another element, there are no intervening elements present. Likewise, it will be understood that when an element such as a layer, region, or substrate is referred to as being "over" or extending "over" another element, it can be directly over or extend directly over the other element or intervening elements may also be present. In contrast, when an element is referred to as being "directly over" or extending "directly over" another element, there are no intervening elements present. It will also be understood that when an element is referred to as being "connected" or "coupled" to another element, it can be directly connected or coupled to the other element or intervening elements may be present. In contrast, when an element is referred to as being "directly connected" or "directly coupled" to another element, there are no intervening elements present.

Relative terms such as "below" or "above" or "upper" or "lower" or "horizontal" or "vertical" may be used herein to describe a relationship of one element, layer, or region to another element, layer, or region as illustrated in the figures. It will be understood that these terms and those discussed previously are intended to encompass different orientations of the device in addition to the orientation depicted in the figures.

The terminology used herein is for the purpose of describing particular embodiments only and is not intended to be limiting of the disclosure. As used herein, the singular forms "a," "an," and "the" are intended to include the plural forms as well, unless the context clearly indicates otherwise. It will be further understood that the terms "comprises," "comprising," "includes," and/or "including" when used herein specify the presence of stated features, integers, steps, operations, elements, and/or components but do not preclude the presence or addition of one or more other features, integers, steps, operations, elements, components, and/or groups thereof.

Unless otherwise defined, all terms (including technical and scientific terms) used herein have the same meaning as commonly understood by one of ordinary skill in the art to which this disclosure belongs. It will be further understood that terms used herein should be interpreted as having a meaning that is consistent with their meaning in the context of this specification and the relevant art and will not be interpreted in an idealized or overly formal sense unless expressly so defined herein.

The present disclosure relates to filter circuitry for use in communication systems that employ multiple antennas. In general, a communication system may have a transmit path and a receive path. The transmit path extends to a first antenna port and is configured to present signals for transmission in a first communication band and a second communication band to the first antenna port for transmission via a first antenna that is coupled to the first antenna port. The receive path extends to a second antenna port and comprises a first multiple passband/multiple stopband (MPMS) filter that provides a plurality of passbands and a plurality of stopbands interleaved with one another, wherein a first stopband and a second stopband of the plurality of stopbands correspond respectively to the first communication band and the second communication band and are separated by a first passband of the plurality of passbands. In one embodiment, the first passband of the plurality of passbands corresponds to a third communication band, and the receive path is configured to receive signals via a second antenna coupled to the second antenna port. Further details provided below.

Today's mobile terminals must communicate using different communication technologies in different bands, which vary significantly in both bandwidth and frequency. To further complicate matters, data rates are ever increasing and the there is a need to transmit and receive over these different bands at the same time. As a result, mobile terminals have very complicated front-end configurations and are starting to employ multiple input multiple output (MIMO) transmission and reception technology, which requires the use of multiple antennas. FIG. **7** is a block diagram of a mobile terminal **42** that incorporates four antennas: a primary antenna A**1**, a secondary antenna A**2**, a tertiary antennae A**3**, and a quaternary antenna A**4**. The mobile terminal **42** generally includes control circuitry **44**, which is associated with a user interface (I/F) **46**, and radio frequency (RF) circuitry **48**. The user interface **46** may include microphones, speakers, keypads, touchscreens, displays, and the like. The RF circuitry **48** may include baseband, transceiver, power amplifier, and switching circuitry, as will be appreciated by those skilled in the art.

In general, signals to be transmitted are provided by the RF circuitry **48** to one or more of the antennas A**1** through A**4**, and signals received by one or more of the antennas A**1** through A**4** are routed to the RF circuitry **48** for demodulation and associated processing. The RF circuitry **48** may be configured to facilitate any number of communications, including first, second, third, fourth, and fifth generation cellular communications, wireless local area network (WLAN) communications, Bluetooth communications, industrial, scientific and medical (ISM) communications, near field communications, and the like. Any of these communications may use MIMO for transmission, reception, or both, depending on the capabilities of the mobile terminal **42** and the systems with which the mobile terminal **42** communicates.

Since mobile terminals **42** are relatively small, the multiple antennas A**1** through A**4** used for MIMO are relatively close to one another. As a result, the antennas A**1** through A**4** may interact with one another, and as a result, modify each other's radiation patterns, which generally alters the antenna's radiation efficiency. With continued reference to FIG. **7**, when the primary antenna A**1** is used for transmission and the tertiary antennae A**3** is used for reception at the same time, the transmission from antennae A**1** may significantly degrade the ability to receive signals via antenna A**3**, given the proximity of antenna A**1** to antenna A**3**. Further, the secondary antenna A**2** and the quaternary antenna A**4** may also be impacted by transmissions from the primary antenna A**1**. As such, there is a need for a cost effective and space efficient technique to resolve, or at least significantly reduce, the impact that one antenna has on another in devices such as the mobile terminal **42** illustrated in FIG. **7**.

With reference to FIG. **8**, a technique for addressing the above issues is described. As illustrated, the RF circuitry **48** is associated with antennas A**1** through A**4**. Each of the antennas is associated with antenna tuning circuitry **50**, **52**, **54**, **56**, respectively. In particular, antenna A**1** is coupled to the RF circuitry **48** through coaxial cable **58**. Antenna A**3** is coupled to the RF circuitry **48** through coaxial cable **60** and a MPMS filter **62**; antenna A**2** is coupled to the RF circuitry **48** through coaxial cable **64** and MPMS filter **66**; and antenna A**4** is coupled to the RF circuitry **48** through coaxial cable **68** and MPMS filter **70**. While MPMS filters **62**, **66**, and **70** are provided for antennas A**3**, A**2**, and A**4**, respectively, alternative embodiments may only employ MPMS filter **62**, given the proximity of antennas A**1** and A**3**. In other embodiments, an appropriately configured MPMS filter (not illustrated) may also be provided in association with antenna A**1**. In short, MPMS filters may be provided for each antenna A**1** through A**4** or any combination thereof. The antenna tuning circuitry **50**, **52**, **54**, **56** are used for tuning impedances associated with the respective antennas A**1**, A**3**, A**2**, A**4**, as those skilled in the art will appreciate.

For the following description, MPMS filter **62** is described in detail; however, MPMS filter **66** and **70** may be similarly or identically configured, depending on the embodiment. Assume that the RF circuitry **48** is configured to transmit RF signals in band X and band Y via antenna A**1** at the same or different times. Further assume that RF circuitry **48** is configured to receive RF signals and bands A, B, and C at the same or different times. Given the proximity of antennas A**1** and A**3**, transmitting in bands X or Y via antenna A**1** would significantly impact the ability of antenna A**3** to receive RF signals in bands A, B, or C, in the absence of MPMS filter **62**. However, adding MPMS filter **62** in close proximity to antenna A**3** significantly reduces the impact that antenna A**1** has on antenna A**3**.

MPMS filter **62** is a specially configured filter that has multiple passbands and multiple stopbands, which are interleaved with one another, as illustrated in FIG. **8**. In this example, passbands are provided for bands A, B, and C and stopbands are provided for at least bands X and Y. A stopband for band X is between passbands for bands A and B, and a stopband for band Y is between passbands for bands B and C. In other words, stopbands are provided for the problematic bands that are transmitted via antenna A**1**, and passbands are provided for the bands to be received via antenna A**3**. The RF circuitry **48** may also transmit signals in bands A, B, or C via antenna A**3**. Providing stopbands for an adjacent antenna's transmission bands and passbands for the selected antenna's receive (and transmission) bands can significantly improve the performance of both antennas. When multiple ones of the MPMS filters **62**, **66**, **70** are employed, the passbands and stopbands may be the same or different amongst the different MPMS filters **62**, **66**, **70**, based on the proximity of the antennas A**1**-A**4** as well as the communication bands used for communications by the mobile terminal **42**.

In certain embodiments, at least two of the stopbands and/or passbands provided by one or more of the MPMS filters **62**, **66**, **70** reside entirely above 2 GHz and have a bandwidth of at least 20 MHz. In other embodiments, at least two of the stopbands and/or passbands reside between 2 GHz and 12 GHz and have a bandwidth of at least 20 MHz, 40 MHz, 50 MHz, or 100 MHz. In select embodiments, at least one of the stopbands or passbands residing between two other stopbands or passbands has a bandwidth of at least 100 MHz, 150 MHz, or 200 MHz. All, or at least certain of, the stopbands may provide attenuation of at least 10 dB, 20 dB, or 30 dB in each of the above embodiments, depending on the configuration of the MPMS filters **62**, **66**, **70**.

With reference to FIG. **9**, the mobile terminal **42** may employ multiple printed circuit boards (PCBs) to implement the necessary electronics for operation. Further, the various antennas A**1**-A**4** may be spread about the mobile terminal **42**. These antennas A**1**-A**4** may be implemented on or in a housing H (illustrated in FIG. **7**) of the mobile terminal **42**, on the various PCBs, or combination thereof. FIG. **9** illustrates a rigid PCB (R-PCB) and a flexible PCB (F-PCB), which are used to implement at least part of the electronics for the mobile terminal **42**. In one embodiment, the rigid PCB R-PCB may be a traditional glass-reinforced multilayer circuit board, wherein the flexible PCB F-PCB is provided by a much thinner, flexible substrate on which traces and components may be formed or mounted. The flexible PCB F-PCB will have a flex factor of at least ten times that of the rigid PCB R-PCB.

As illustrated, the control circuitry **44** and the RF circuitry **48** are implemented in whole or in part on the rigid PCB R-PCB while the MPMS filters **62**, **66**, **70** and the antenna tuning circuitry **50**, **52**, **54**, **56** are implemented on the flexible PCB F-PCB. The coaxial cables **58**, **60**, **64**, **68** connect the rigid PCB R-PCB and the flexible PCB F-PCB such that the transmit/receive paths that extend between the RF circuitry **48** and the respective antennas A**1**, A**2**, A**3**, and A**4** are provided by the combination of the rigid PCB R-PCB, the flexible PCB F-PCB, and the coaxial cables **58**, **60**, **64**, **68**. These transmit/receive paths extend to corresponding antenna ports AP**1**, AP**2**, AP**3**, AP**4** of the flexible PCB F-PCB. The antennas A**1**, A**2**, A**3**, and A**4** are connected to the antenna ports AP**1**, AP**2**, AP**3**, AP**4**, respectively, through cables, traces, and/or the like.

With reference to FIG. **10**, a low noise amplifier (LNA) **72** may be provided between the MPMS filter **62** and the coaxial cable **60** to amplify the filtered receive signals prior to the coaxial cable **60**. The LNA **72** may be provided along with the MPMS filter **62** on the flexible PCB F-PCB, wherein the RF circuitry **48** is provided on the rigid PCB R-PCB, and the coaxial cable **60** connects the flexible PCB F-PCB and the rigid PCB R-PCB. The antenna tuning circuitry **50**, **52** may also be provided on the flexible PCB F-PCB.

The following provides various filters that employ acoustic resonators and are capable of providing a filter response that includes multiple passbands and stopbands. Some basics regarding the theory of operation are provided prior to describing the specific configurations, which provide the desired filter responses.

Turning now to FIG. **11**, a series resonator B**1** is shown coupled between an input node I/P and an output node O/P. The series resonator B**1** has a series resonance frequency F_s and inherent capacitance, which generally limits the bandwidth of filters that employ the series resonator B**1**. In the case of a Bulk Acoustic Wave (BAW) resonator, the capacitance of the series resonator B**1** is primarily caused by its inherent structure, which looks and acts like a capacitor in part because the series resonator includes the top and bottom electrodes **20**, **22** (FIG. **1**) that are separated by a dielectric piezoelectric layer **18**. While BAW resonators are the focus of the example, other types of acoustic resonators, such as Surface Acoustic Wave (SAW) resonators, are equally applicable.

A compensation circuit **74** is coupled in parallel with the series resonator B**1** and functions to compensate for some of the capacitance presented by the series resonator B**1**. The compensation circuit **74** includes two negatively coupled inductors L**1**, L**2** and a shunt resonator B**2**. The inductors L**1**, L**2** are coupled in series between the input node I/P and the output node O/P, wherein a common node CN is provided between the inductors L**1**, L**2**. The inductors L**1**, L**2** are magnetically coupled by a coupling factor K, wherein the dots illustrated in association with the inductors L**1**, L**2** indicate that the magnetic coupling is negative. As such, the inductors L**1**, L**2** are connected in electrical series and negatively coupled from a magnetic coupling perspective. As defined herein, two (or more) series-connected inductors that are negatively coupled from a magnetic perspective are inductors that are

- connected in electrical series; and
- the mutual inductance between the two inductors functions to decrease the total inductance of the two (or more) inductors.

The shunt resonator B**2** is coupled between the common node CN and ground, or other fixed voltage node.

To compensate for at least some of the capacitance of the series resonator B**1**, the compensation circuit **74** presents itself as a negative capacitance within certain frequency ranges when coupled in parallel with the series resonator B**1**. Since capacitances in parallel are additive, providing a negative capacitance in parallel with the (positive) capacitance of the series resonator B**1** effectively reduces the capacitance of the series resonator B**1**. With the compensation circuit **74**, the series resonator B**1** can actually function as a filter (instead of just a resonator) and provide a passband, albeit a fairly narrow passband, instead of a more traditional resonator response (solid line of FIG. **2**). FIG. **12** graphically illustrates the frequency responses of the series resonator B**1** (inside the block referenced B**1**), the compensation circuit **74** (inside the block referenced **74**), and the overall circuit in which the compensation circuit **74** is placed in parallel with the series resonator B**1**. As illustrated, the overall circuit provides a relatively narrow passband. Further detail on this particular circuit topology can be found in the co-assigned U.S. patent application Ser. No. 15/004,084, filed Jan. 22, 2016, now U.S. Pat. No. 9,837,984, and titled RF LADDER FILTER WITH SIMPLIFIED ACOUSTIC RF RESONATOR PARALLEL CAPACITANCE COMPENSATION, and U.S. patent application Ser. No. 14/757,651, filed Dec. 23, 2015, and titled SIMPLIFIED ACOUSTIC RF RESONATOR PARALLEL CAPACITANCE COMPENSATION, the disclosures of which are incorporated herein by reference in their entireties.

While beneficial in many applications, the narrow passband of the circuit topology of FIG. **11** has its limitations. With the challenges of modern day communication systems, wider passbands and the ability to provide multiple passbands within a given system are needed. Fortunately, applicants have discovered that certain modifications to this topology provide significant and truly unexpected increases in passband bandwidths and, in certain instances, the ability to generate multiple passbands of the same or varying bandwidths in an efficient and effective manner.

With reference to FIG. **13**, a modified circuit topology is illustrated wherein the circuit topology of FIG. **11** is modified to include an additional shunt resonator B**3**, which is coupled between the common node CN and ground. As such, a new compensation circuit **76** is created that includes the negatively coupled inductors L**1** and L**2**, which have a coupling coefficient K, and at least two shunt resonators B**2**, B**3**. The compensation circuit **76** is coupled in parallel with the series resonator B**1**. When the series resonance frequencies F_s of the shunt resonators B**2**, B**3** are different from one another, unexpectedly wide bandwidths passbands are achievable while maintaining a very flat passbands, steep skirts, and excellent cancellation of signals outside of the passbands.

FIG. **14** graphically illustrates the frequency responses of the series resonator B**1** (inside the block referenced B**1**), the compensation circuit **76** (inside the block referenced **76**), and the overall circuit in which the compensation circuit **76** is placed in parallel with the series resonator B**1**. As illustrated, the overall circuit with the compensation circuit **76** provides a much wider passband (FIG. **10**) than the overall circuitry with the compensation circuit **74** (FIG. **12**).

While FIGS. **12** and **14** are graphical representations, FIG. **15** is an actual comparison of the frequency response of the overall circuit using the different compensation circuits **74**, **76**, wherein the overall circuit using the compensation circuit **76** provides a significantly wider and better formed passband (solid line) than the overall circuit using the compensation circuit **74** (dashed line).

As illustrated in FIG. **16**, the concepts described herein not only contemplate the use of multiple shunt resonators B**2**, B**3**, which are coupled between the common node CN and ground, but also multiple series resonators, such as series resonators B**1** and B**4**, which are coupled in parallel with one another between the input node VP and the output node O/P. The series resonance frequencies F_s of the series resonators B**1**, B**4** are different from one another, and the series resonance frequencies F_s of the shunt resonators B**2**, B**3** are also different from one another and different from those of the series resonators. While only two series resonators B**1**, B**4** and two shunt resonators B**2**, B**3** are illustrated, any number of these resonators may be employed depending on the application and the desired characteristics of the overall frequency response of the circuit in which these resonators and associated compensation circuits **76** are employed. While the theory of operation is described further subsequently, FIGS. **17** and **18** illustrate just two of the many possibilities.

For FIG. **17**, there are two series resonators B**1**, B**4** and two shunt resonators B**2**, B**3**, with different and relatively dispersed series resonance frequencies F_s. FIG. **17** graphically illustrates the frequency responses of the combination of the two series resonators B**1**, B**4** (inside the block referenced BX), the compensation circuit **76** with two shunt resonators B**2**, B**3** (inside the block referenced **76**), and the overall circuit in which the compensation circuit **76** is placed in parallel with the series resonator B**1**. As illustrated, the overall circuit in this configuration has the potential to provide a passband that is even wider than that for the embodiment of FIGS. **13** and **14**. For example, passbands of greater than 100 MHz, 150 MHz, 175 MHz, and 200 MHz are contemplated at frequencies at or above 1.5 GHz, 1.75 GHz, and 2 GHz.

In other words, center-frequency-to-bandwidth ratios (fc/BW*100) of 3.5% to 9%, 12%, or greater are possible, wherein fc is the center frequency of the passband and BW is the bandwidth of the passband. If multiple passbands are provided, BW may encompass all of the provided passbands. Further, when multiple passbands are provided, the passbands may have the same or different bandwidths or center-frequency-to-bandwidth ratios. For example, one passband may have a relatively large center-frequency-to-bandwidth ratio, such as 12%, and a second passband may have a relatively small center-frequency-to-bandwidth ratio, such as 2%. Alternatively, multiple ones of the passbands may have a bandwidth of 100 MHz, or multiple ones of the passbands may have generally the same center-frequency-to-bandwidth ratios. In the latter case, the bandwidths of the passbands may inherently be different from one another, even though the center-frequency-to-bandwidth ratios are the same.

For FIG. **18**, there are four series resonators, which are coupled in parallel with one another (not shown), and two shunt resonators (not shown) with different and more widely dispersed series resonance frequencies F_s. FIG. **18** graphically illustrates the frequency responses of the combination of the four series resonators (inside the block referenced BX), the compensation circuit **76** with two shunt resonators B**2**, B**3** (inside the block referenced **76**), and the overall circuit in which the compensation circuit **76** is placed in parallel with the series resonator B**1**. As illustrated, the overall circuit in this configuration provides multiple passbands, which are separated by a stopband. In this embodiment, two passbands are provided; however, the number of passbands may exceed two. The number of passbands in the bandwidth of each of the passbands is a function of the number of shunt and series resonators B**1**-B**4** and the series resonance frequencies F_s thereof.

The theory of the compensation circuit **76** follows and is described in association with FIGS. **19**A through **19**D and FIG. **20**. With reference to FIG. **19**A, assume the compensation circuit **76** includes the two negatively coupled inductors L**1**, L**2**, which have an inductance value L, and two or more shunt resonators BY, which have an overall shunt impedance Zres presented between the common node CN and ground. While the inductance values L of the negatively coupled inductors L**1**, L**2** are described as being the same, these values may differ depending on the application. Also assume that the one or more series resonators BX present an overall series impedance ZS.

As shown in FIG. **19**B, the two negatively coupled and series-connected inductors L**1**, L**2** (without Zres) can be modeled as a T-network of three inductors L**3**, L**4**, and L**5**, wherein series inductors L**3** and L**4** are connected in series and have a value of L(1+K), and shunt inductor L**5** has a value of −L*K, where K is a coupling factor between the negatively coupled inductors L**1**, L**2**. Notably, the coupling factor K is a positive number between 0 and 1. Based on this model, the overall impedance of the compensation circuit **76** is modeled as illustrated in FIG. **19**C, wherein the shunt impedance Zres is coupled between the shunt inductor L**5** and ground. The resulting T-network, as illustrated in FIG. **19**C, can be transformed into an equivalent π (pi) network, as illustrated in FIG. **19**D.

The **7** network of FIG. **19**D can be broken into a series impedance ZA and two shunt equivalent impedances ZB. The series equivalent impedance ZA is represented by two series inductances of value L*(1+K), where K>0, and a special "inversion" impedance Zinv. The inversion impedance Zinv is equal to $[L(1+K)\omega]^2/[Zres-jLK\omega]$, where $\omega=2\pi f$ and f is the frequency. As such, the series equivalent impedance ZA equals j*2*L(1+K)ω+Zinv and is coupled between the input node I/O and the output node O/P. Each of the two shunt equivalent impedances ZB is represented by an inductor of value L(1−K) in series with two overall shunt impedances Zres.

Notably, the series equivalent impedance ZA has a negative capacitor behavior at certain frequencies at which broadband cancellation is desired and has series resonance at multiple frequencies. In general, the series equivalent impedance ZA has a multiple bandpass-bandstop characteristic, in that the series equivalent impedance ZA will pass some frequencies and stop others. When the series equivalent impedance ZA is placed in parallel with the series impedance Zs of the series resonators BX, which can also have a multiple bandpass-bandstop characteristic, a broadband filter or a filter with multiple passbands may be created.

FIG. **20** illustrates the series impedance ZS of the series resonators BX in parallel with the series equivalent impedance ZA of the compensation circuit **76**. The overall series impedance ZAs represents the series impedance ZS in parallel with the series equivalent impedance ZA. The two shunt impedances ZB are respectively coupled between the input port I/P and ground and the output port O/P and ground. The primary focus for the following discussion relates to the series equivalent impedance ZA and its impact on the series impedance ZS when the series equivalent impedance ZA is placed in parallel with the series impedance ZS.

As noted previously, the series equivalent impedance ZA provides two primary functions. The first provides a negative capacitive behavior, and the second provides one or more additional series resonances between the input node I/P and the output node O/P. These additional series resonances are provided through the series equivalent impedance ZA and are in addition to any series resonances that are provided through the series impedance ZS of the series resonators BX. To help explain the benefits and concept of the negative capacitive behavior provided by the series equivalent impedance ZA, normal capacitive behavior is illustrated in association with the overall shunt impedance Zres, which is provided by the shunt resonators BY. FIG. **21** graphs the absolute (magnitude) and imaginary components of the overall shunt impedance Zres, which is formed by two shunt resonators BY, which are coupled in parallel with one another.

The series resonance frequency F_s for each of the two shunt resonators BY occurs when the absolute impedance (abs(Zres)) is at or near zero. Since there are two shunt resonators BY, the absolute impedance (abs(Zres)) is at or near zero at two frequencies, and as such, there are two series resonance frequencies F_s. The parallel resonance frequencies F_p occur when the imaginary component (imag (Zres)) peaks. Again, since there are two shunt resonators BY, there are two series resonance frequencies F_s provided by the overall shunt impedance Zres.

Whenever the imaginary component (imag(Zres)) of the overall shunt impedance Zres is less than zero, the overall shunt impedance Zres has a capacitive behavior. The capacitive behavior is characterized in that the reactance of the overall shunt impedance Zres is negative and decreases as frequency increases, which is consistent with capacitive reactance, which is represented by 1/jωC. The graph of FIG. **21** identifies three regions within the impedance response of the overall shunt impedance Zres that exhibit capacitive behavior.

Turning now to FIG. **22**, the series equivalent impedance ZA is illustrated over the same frequency range as that of the overall shunt impedance Zres, which was illustrated in FIG. **21**. The series equivalent impedance ZA has two series resonance frequencies F_s, which occur when the absolute impedance (abs(ZA)) is at or near zero. The two series resonance frequencies F_s for the series equivalent impedance ZA are different from each other and slightly different from those for the overall shunt impedance Zres. Further, the number of series resonance frequencies F_s generally corresponds to the number of shunt resonators BY in the compensation circuit **76**, assuming the series resonance frequencies F_s are different from one another.

Interestingly, the imaginary component (imag(ZA)) of the series equivalent impedance ZA is somewhat inverted with respect to that of the overall shunt impedance Zres. Further, the imaginary component (imag(ZA)) of the series equivalent impedance ZA has a predominantly positive reactance. During the portions at which the imaginary component (imag(ZA)) is positive, the reactance of the series equivalent impedance ZA again decreases as frequency increases, which is indicative of capacitive behavior. However, the reactance is positive, whereas traditional capacitive behavior would present a negative reactance. This phenomenon is referred to as negative capacitive behavior. Those portions of the imaginary component (imag(ZA)) of the series equivalent impedance ZA that are positive and thus exhibit negative capacitive behavior are highlighted in the graph of FIG. **22**.

The negative capacitive behavior of the series equivalent impedance ZA for the compensation circuit **76** is important, because when the series equivalent impedance ZA is placed in parallel with the series impedance ZS, the effective capacitance of the overall circuit is reduced. Reducing the effective capacitance of the overall circuit shifts the parallel resonance frequency F_p of the series impedance ZS higher in the frequency range, which is described subsequently, and significantly increases the available bandwidth for passbands while providing excellent out-of-band rejection.

An example of the benefit is illustrated in FIGS. **23**A and **23**B. The thicker solid line, which is labeled abs(VG), represents the frequency response of the overall circuit illustrated in FIG. **16**, wherein there are two series resonators BX and two shunt resonators BY in the compensation circuit **76**. The frequency response has two well-defined passbands, which are separated by a stopband. The frequency response abs(VG) of the overall circuit generally corresponds to the inverse of the overall series impedance ZAs, which again represents the series impedance Zs in parallel with the series equivalent impedance ZA, as provided in FIG. **20**.

Notably, the parallel resonance frequencies F_p(ZS) of the series impedance ZS, in isolation, fall in the middle of the passbands of frequency response abs(VG) of the overall circuit. If the parallel resonance frequencies F_p(ZS) of the series impedance ZS remained at these locations, the passbands would be severely affected. However, the negative capacitive behavior of the series equivalent impedance ZA functions to shift these parallel resonance frequencies F_p(ZS) of the series impedance ZS to a higher frequency and, in this instance, above the respective passbands. This is manifested in the resulting overall series impedance ZAs, in which the only parallel resonance frequencies F_p(ZAs) occur above and outside of the respective passbands. An additional benefit to having the parallel resonance frequencies F_p(ZAs) occur outside of the respective passbands is the additional cancellation of frequencies outside of the passbands. Plus, the overall series impedance ZAs is lower than the series impedance ZS within the respective passbands.

A further contributor to the exemplary frequency response abs(VG) of the overall circuit is the presence of the additional series resonance frequencies F_s, which are provided through the series equivalent impedance ZA. These series resonance frequencies F_s are offset from each other and from those provided through the series impedance ZS. The series resonance frequencies F_s for the series equivalent impedance ZA in the series impedance ZS occur when the magnitudes of the respective impedances approach zero. The practical results are wider passbands, steeper skirts for the passbands, and greater rejection outside of the passbands, as evidenced by the frequency response abs(VG) of the overall circuit.

Turning now to FIG. **24**, another embodiment is provided wherein the compensation circuit **76** is placed in parallel with one or more series resonators BX. In this embodiment, shunt resonator B**2** is permanently coupled between the common node CN and ground. Shunt resonators B**3**, B**5**, and B**6** can be selectively coupled between the common node CN and ground via respective switches S**1**, S**2**, and S**3**. By using control circuitry (not shown) to selectively switch the various shunt resonators B**3**, B**5**, and B**6** into and out of the compensation circuit **76**, the passbands and stopbands provided by the overall circuit can be dynamically adjusted for different modes of operation. Again, the series resonance frequencies F_s of the shunt resonators B**2**, B**3**, B**5**, and B**6** will generally differ from one another. Resistors R**1** and R**2** are illustrated and may be coupled between the input node VP and ground and the output node O/P and ground, respectively. In one embodiment, the series resonance frequency F_s of the series equivalent impedance ZA is greater than the series resonance frequency F_s of the series impedance ZS. The series resonance frequency F_s of at least one of the shunt resonators B**2**, B**3**, B**5**, and B**6** is greater than the series resonance frequency F_s of the series impedance ZS. As with any of these embodiments, the number of shunt resonators BY and series resonators BX may vary from embodiment to embodiment. The number illustrated is merely for illustrative purposes.

FIG. **25** illustrates yet another embodiment, which is similar to that illustrated in FIG. **24**. The difference is that the series resonators BX are replaced with a lumped series L-C circuit, which is formed from a series capacitors CS and a series inductor LS that are coupled in series between the input node I/P and the output node O/P.

FIG. **26** illustrates an embodiment similar to that of FIG. **13**, except that at least one inductor L**6** is coupled in series with shunt resonator B**3**. As such, shunt resonator B**2** is coupled between the common node CN and ground without a series inductor, and shunt resonator B**2** and inductor L**6** are coupled in series with one another and between the common node CN and ground. In one embodiment, the series resonance frequency F_s of the series equivalent impedance ZA is greater than the lowest series resonance frequency F_s of the series impedance ZS. FIG. **27** illustrates a further modification to the embodiment of FIG. **26**, wherein an inductor L**7** is placed in series with the series resonators BX, such that the inductor L**7** and the series resonators BX are coupled in series with one another between the input node I/P and the output node O/P.

With reference to FIG. **28**, a more complex filter arrangement is illustrated. In particular, resonators B**7**, B**8**, and B**9** are coupled in series between the input node I/P and the output node O/P. The compensation circuits **76** of FIG. **26** are coupled across resonator B**7** and resonator B**9**. The resonators B**7**, B**8**, and B**9** may each represent a single acoustic resonator or multiple acoustic resonators in parallel.

FIG. **29** illustrates yet another embodiment wherein resonators B**10** and B**11** are coupled in series between the input node I/P and the output node O/P. An inductor L**8** is coupled between node N**1** and ground. The compensation circuit **76** is coupled across both of the resonators B**10** and B**11**. Accordingly, the compensation circuit **76** may be coupled across one or more acoustic resonators along the series path that extends between the input node I/P and the output node O/P. FIG. **30** illustrates a modification to the embodiment of FIG. **29**, wherein the inductor L**8** is replaced with a shunt resonator B**12**.

Turning now to FIG. **31**, an alternative MPMS filter **80** is illustrated. The MPMS filter **80** extends between an antenna port AP**3**, which is coupled to the antenna A**3**, and an RF port RFP, which is coupled to the RF circuitry **48**. The MPMS filter **80** includes a series path that has parallel resonators BX**1**, which have different resonance frequencies. A first shunt path between the antenna port AP**3** and ground includes parallel resonators BX**2** in series with a first inductor L**3**. A second shunt path provides for an inductor L**4** between the RF port RFP and ground. The inductors L**3** and L**4** are negatively coupled and have a coupling factor k, which is a positive number between 0 and 1. The parallel resonators BX**2** have resonance frequencies which are different from one another, and depending on the embodiment, different from the resonance frequencies of the parallel resonators BX**1**. With the appropriate selection of resonance frequencies, the MPMS filter **80** may provide multiple stopbands and passbands, which are appropriately located in frequency and have the appropriate bandwidths. As with the above embodiments, the bandwidths for the various stopbands and passbands may be the same or different, depending on the application.

Turning now to FIG. **32**, yet another alternative MPMS filter **82** is illustrated. The MPMS filter **82** again extends between an antenna port AP**3**, which is coupled to the antenna A**3** and an RF port RFP, which is coupled to the RF circuitry **48**. The MPMS filter **82** includes a series path that has parallel resonators BX**1**, which have different resonance frequencies. A first path between the antenna port AP**3** and a node N**1** includes parallel resonators BX**2** in series with a first inductor L**3**. A second path provides for an inductor L**4** between the RF port RFP and node N**1**. The inductors L**3** and L**4** are negatively coupled and have a coupling factor k, which is a positive number between 0 and 1. The parallel resonators BX**2** have resonance frequencies which are different from one another, and depending on the embodiment, different from the resonance frequencies of the parallel resonators BX**1**. Further, parallel resonators BY are coupled between node N**1** and ground. The parallel resonators BY have resonance frequencies which are different from one another, and depending on the embodiment, different from the resonance frequencies of the parallel resonators BX**1** and BX**2**. With the appropriate selection of resonance frequencies, the MPMS filter **80** may provide multiple stopbands and passbands, which are appropriately located in frequency and have the appropriate bandwidths. As with the above embodiments, the bandwidths for the various stopbands and passbands may be the same or different, depending on the application.

While the concepts disclosed herein were described in association with a mobile terminal, these concepts are applicable to any type of communication device that employs wireless communications. Those skilled in the art will recognize numerous modifications and other embodiments that incorporate the concepts described herein. These modifications and embodiments are considered to be within scope of the teachings provided herein and the claims that follow.

What is claimed is:

1. An apparatus comprising:

a transmit path extending to a first antenna port and configured to present signals for transmission in a first communication band and a second communication band to the first antenna port for transmission via a first antenna that is coupled to the first antenna port; and

a receive path extending to a second antenna port and comprising a first multiple passband/multiple stopband (MPMS) filter that provides a plurality of passbands and a plurality of stopbands interleaved with one another, wherein a first stopband and a second stopband of the plurality of stopbands correspond respectively to the first communication band and the second communication band and are separated by a first passband of the plurality of passbands.

2. The apparatus of claim **1** wherein the first passband of the plurality of passbands corresponds to a third communication band, and the receive path is configured to receive signals received via a second antenna coupled to the second antenna port.

3. The apparatus of claim **2** further comprising radio frequency circuitry configured to generate the signals for transmission as well as demodulate the signals received via the second antenna.

4. The apparatus of claim **3** further comprising a first coaxial cable electrically coupled between the radio frequency circuitry and the transmit path, and a second coaxial cable electrically coupled between the radio frequency circuitry and the receive path.

5. The apparatus of claim **4** further comprising a rigid printed circuit board on which the radio frequency circuitry is provided and a flexible printed circuit board on which the receive path, including the first MPMS filter, is provided.

6. The apparatus of claim **5** wherein the transmit path is provided on the flexible printed circuit board and the first coaxial cable extends between the rigid printed circuit board and the flexible printed circuit board.

7. The apparatus of claim **6** wherein the first antenna is coupled to a first antenna port of the flexible printed circuit board, and the second antenna is coupled to a second antenna port of the flexible printed circuit board.

8. The apparatus of claim **3** wherein the radio frequency circuitry is further configured to receive signals being received via the first antenna and demodulate the signals received via the second antenna and the signals received via the first antenna together using multiple input multiple output processing.

9. The apparatus of claim **1** further comprising a low noise amplifier in the receive path and configured to amplify signals received via a second antenna coupled to the second antenna port.

10. The apparatus of claim **1** wherein the first stopband and second stopband reside entirely above 2 GHz.

11. The apparatus of claim **10** wherein a bandwidth of the first passband is at least 20 MHz and a bandwidth of the second stopband is at least 20 MHz.

12. The apparatus of claim **10** wherein a bandwidth of the first passband is at least 100 MHz.

13. The apparatus of claim **12** wherein a bandwidth of the second stopband is at least 20 MHz.

14. The apparatus of claim **1** wherein the first MPMS filter comprises:

a first node and a second node;

a first plurality of acoustic resonators coupled in parallel between the first node and the second node;

a first shunt path comprising a series combination of a second plurality of acoustic resonators and a first inductor coupled between the first node and a fixed voltage node; and

a second shunt path comprising a second inductor coupled between the second node and the fixed voltage node, wherein the first inductor and the second inductor are negatively coupled with one another.

15. The apparatus of claim **1** wherein the first MPMS filter comprises:

a first node and a second node;

a first plurality of acoustic resonators coupled in parallel between the first node and the second node;

a first shunt path comprising a series combination of a second plurality of acoustic resonators and a first inductor coupled between the first node and a fixed voltage node; and

a second shunt path comprising a second inductor coupled between the second node and the fixed voltage node, wherein the first inductor and the second inductor are negatively coupled with one another.

16. The apparatus of claim **1** wherein the first MPMS filter comprises:

a first node and a second node;

a first plurality of acoustic resonators coupled in parallel between the first node and the second node;

a first shunt path comprising a series combination of a second plurality of acoustic resonators and a first inductor coupled between the first node and a common node;

a second shunt path comprising a second inductor coupled between the second node and the common node, wherein the first inductor and the second inductor are negatively coupled with one another; and

a third plurality of acoustic resonators coupled between the common node and a fixed voltage node.

17. The apparatus of claim **1** wherein the first MPMS filter comprises:

a first node and a second node;

at least one series acoustic resonator coupled between the first node and the second node, wherein at least one main series resonance is provided between the first node and the second node at a main resonance frequency through the at least one series acoustic resonator; and

a compensation circuit comprising:

a first inductor and a second inductor coupled in series between the first node and the second node, wherein the first inductor and the second inductor are negatively coupled with one another and a common node is provided between the first inductor and the second inductor;

a first shunt acoustic resonator coupled between the common node and a fixed voltage node; and

a second shunt acoustic resonator coupled between the common node and the fixed voltage node, wherein a first series resonance at a first resonance frequency and a second series resonance at a second resonance frequency, which is different from the first resonance frequency and the main resonance frequency, are provided between the first node and the second node through the compensation circuit.

18. The apparatus of claim **17** wherein at least one of the first resonance frequency and the second resonance frequency is greater than the main resonance frequency.

19. The apparatus of claim **17** wherein the first resonance frequency is less than the main resonance frequency, and the second resonance frequency is greater than the main resonance frequency.

20. The apparatus of claim **17** wherein the at least one series acoustic resonator comprises a plurality of acoustic resonators that are coupled in parallel with one another, and each of the plurality of acoustic resonators has a different series resonance frequency.

21. The apparatus of claim **20** wherein the compensation circuit comprises at least one additional shunt acoustic resonator coupled between the common node and the fixed voltage node.

22. The apparatus of claim **17** wherein the at least one series acoustic resonator, the first shunt acoustic resonator, and the second shunt acoustic resonator are at least one of a group consisting of bulk acoustic wave (BAW) resonator and a surface acoustic wave (SAW).

23. The apparatus of claim **1** wherein first MPMS filter comprises at least one acoustic resonator of a group consisting of bulk acoustic wave (BAW) resonator and a surface acoustic wave (SAW).

* * * * *

UNITED STATES PATENT AND TRADEMARK OFFICE

CERTIFICATE OF CORRECTION

PATENT NO. : 10,141,644 B2
APPLICATION NO. : 15/490381
DATED : November 27, 2018
INVENTOR(S) : Nadim Khlat and Robert Aigner

It is certified that error appears in the above-identified patent and that said Letters Patent is hereby corrected as shown below:

In the Specification

In Column 12, Line 45, replace “input node VP” with --input node I/P--.

In Column 14, Line 1, replace “The 7 network” with --The π network--.

In Column 16, Line 36, replace “VP” with --I/P--.

Signed and Sealed this
Twenty-second Day of January, 2019

Andrei Iancu
Director of the United States Patent and Trademark Office